GEORGIA INSTITUTE OF TECHNOLOGY — OFFICE OF CONTRACT ADMINISTRATION

PROJECT ADMINISTRATION DATA SHEET

[x] ORIGINAL [ ] REVISION NO. ______

Project No. A-3517 GTRI/~~GIT~~ DATE 4/18/83

Project Director: N. Walter Cox ~~School~~/Lab EML/PSD

Sponsor: International Telephone & Telegraph Corporation

Type Agreement: Standard Research Project Agreement & P. O. No. X1173

Award Period: From 4/7/83 To 12/30/83 (Performance) 12/30/83 (Reports)

Sponsor Amount: Total Estimated: $ 72,464* Funded: $ 72,464

Cost Sharing Amount: $ ______ Cost Sharing No: ______

Title: Millimeter Wave Receiver Study

ADMINISTRATIVE DATA

OCA Contact Frank Huff

1) Sponsor Technical Contact:

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Defense Priority Rating: NA

Military Security Classification: ______
(or) Company/Industrial Proprietary: See below*

RESTRICTIONS

See Attached NA Supplemental Information Sheet for Additional Requirements.

Travel: Foreign travel must have prior approval — Contact OCA in each case. Domestic travel requires sponsor approval where total will exceed greater of $500 or 125% of approved proposal budget category.

Equipment: Title vests with None proposed

COMMENTS:

*Contract amount of $72,464 includes $2,500 patent & data rights fee paid by sponsor. Article 12 enables sponsor to restrict publication during the first year after disclosure of intent to publish.

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EORGIA INSTITUTE OF TECHNOLOGY OFFICE OF CONTRACT ADMINISTRATION

SPONSORED PROJECT TERMINATION/CLOSEOUT SHEET

5R327

Date 3/2/84

oject No. A-3517 ~~School/~~Lab EML

cludes Subproject No.(s) ----

oject Director(s) Cox/Rucker GTRI / ~~GIT~~

onsor ITT-Electro-Optical Products Division

tle "Millimeter Wave Receiver Study"

fective Completion Date: 2/20/84 (Performance) 2/20/84 (Reports)

NOTE: Completion date is verbal extension from 12/30/83.

ant/Contract Closeout Actions Remaining:

- [ ] None
- [x] Final Invoice ~~or Final Fiscal Report~~
- [ ] Closing Documents
- [ ] Final Report of Inventions
- [ ] Govt. Property Inventory & Related Certificate
- [ ] Classified Material Certificate
- [ ] Other ________

ntinues Project No. ________ Continued by Project No. ________

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Del No. 1

Monthly Status Report 1

MILLIMETER WAVE RECEIVER STUDY

Contract period covered
7 April 1983 through 30 April 1983
P.O. No. X1173

A-3517

Submitted to
ITT
Electro-Optical Products Division
7635 Plantation Road
Roanoke, Virginia 24019

by
Physical Sciences Division
Electromagnetics Laboratory
Engineering Experiment Station
Georgia Institute of Technology
Atlanta, Georgia 30332

Contracting through
Georgia Tech Research Institute
Georgia Institute of Technology
Atlanta, Georgia 30332

May 12, 1983

The technical effort on this program has not begun due to difficulties in shifting staff from other assignments. Plans are to initiate the millimeter wave receiver study on approximately 19 May 1983. No expenditures were made to the program during this reporting period.

Del No. 2

Monthly Status Report 2

MILLIMETER WAVE RECEIVER STUDY

Contract period covered
1 May 1983 through 31 May 1983
P.O. No. X1173

A-3517
Submitted to
ITT
Electro-Optical Products Division
7635 Plantation Road
Roanoke, Virginia 24019

by
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Contracting through
Georgia Tech Research Institute
Georgia Institute of Technology
Atlanta, Georgia 30332

June 15, 1983

WORK PERFORMED DURING THIS PERIOD

A kickoff meeting was held with the engineers and scientists that have been assigned to this project. A general plan of attach was established for this study program directed towards the development of a monolithic millimeter wave receiver, and task assignments were made.

The early stages of this program will involve an in-depth literature search to ascertain what has been done with various types of transmission media and receiver configurations and to evaluate the results. Trade-off studies are to be performed based on receiver block diagrams currently being developed. Device performance, size, and choice of transmission media are some of the areas which will be studied in these early stages on this program. Innovative ideas and techniques will be emphasized and evaluated.

Georgia Tech's past experience in millimeter wave mixer and receiver development has emphasized hardware for fielded systems. The mixers developed have been innovative in nature and include subharmonically and fundamentally pumped balanced mixers, single-ended mixers, and orthogonal waveguide balanced mixers. The tranmission media used for these devices include coaxial line, microstrip, enclosed microstrip, suspended substrate stripline, rectangular waveguide, ridged waveguide and circular waveguide. Hybrid combinations of these media with appropriate transitions have yielded excellent results from 30 to 300 GHz. The types of diodes developed at Georgia Tech and used for these mixers include both whisker-contacted Schottky diode arrays as well as planar Schottky barrier diodes. This experience base will be used to compare and evaluate the receiver alternatives.

WORK TO BE PERFORMED DURING THE NEXT PERIOD

The literature search will begin and development of receiver block diagrams and trade-off studies will continue.

Del. No. 3

Monthly Status Report 3

MILLIMETER WAVE RECEIVER STUDY

Contract period covered
1 June 1983 through 30 June 1983
P.O. No. X1173

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ITT
Electro-Optical Products Division
7635 Plantation Road
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Atlanta, Georgia 30332

July 15, 1983

GENERAL

During June, work has involved start up of the literature search and generation of tentative receiver approaches for trade-off studies. Neither of these tasks is complete at present, but an increased level of effort on both tasks is expected to produce significant results before August 30.

LITERATURE SEARCH/STUDY

The literature search is not being restricted to mixer/receiver approaches only. Instead, we are also considering more fundamental items such as:

- coupling means between various transmission lines,
- FET doubling oscillators,
- dielectric stabilizers,
- circuit losses etc.

Determining which transmission line type (suspended substrate, microstrip, fin-line ----) is likely to be optimum for each circuit function is not a trivial task. One is confronted with an abundance of options and the problem is to try to predict the most viable receiver approach that will ultimately emerge. This approach is necessary to avoid the temptation to proceed with an approach already "successfully" demonstrated by either Georgia Tech or others. The literature search/study will form a continuing portion of the effort.

POTENTIAL APPROACHES

Several generic mixer types are being considered. They are:

- Single ended, using either one mixer diode or two.
- Balanced fundamental, using either
  - a) a hybrid circuit (waveguide/microstrip)

b) orthogonal waveguides (crossbar) or

c) orthogonal fin-line/microstrip geometry.

- Balanced subharmonic, using either

  a) one-half frequency L.O. or

  b) one-fourth frequency L.O.

For each of these, L.O. device types, means of frequency stabilization, filter media and fabrication approach are being compared. Recommended approaches will be prepared and reported by August 30.

WORK FOR NEXT PERIOD

The main objective of the next two work periods will be to conduct the literature assessment and study of approaches to a bottom line "preferred approach" for detailed design.

Del No. 4

Monthly Status Report 4

MILLIMETER WAVE RECEIVER STUDY

Contract period covered
1 July 1983 through 31 July 1983
P. O. No. X1173

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Submitted to
ITT
Electro-Optical Products Division
7635 Plantation Road
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Georgia Tech Research Institute
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Atlanta, Georgia 30332

August 26, 1983

General

During July, this work has received heavy emphasis. Both the literature search and other work are proceeding rapidly. Heavy emphasis will continue in August.

It is anticipated that all schedule items planned for review near the end of August will have received adequate attention by that time to allow initial component development to begin.

We have reached a tentative conclusion that a modified suspended substrate stripline medium is probably optimum for a practical receiver approach at 35 GHz. Detailed design alternates are being prepared based on this conclusion. Rationale for this choice will be presented at the design review to be held in early September. One of the chief problems with suspended substrate approaches is the tendency to break substrates when mounting. Various metallizations applied tend to cause substrate warping. When combined with the small size and close tolerance mounting grooves this warpage can be quite troublesome. We have devised two unique approaches toward solving this problem and reducing the need for close mounting tolerance as well. Both will be reported for discussion in September.

N. W. Cox, C. T. Rucker and R. E. Forsythe plan to visit ITT in early September. A follow-on proposal will also be presented at that time.

WORK FOR NEXT PERIOD

The further definition of the preferred approach will be the main area treated during August.

Del No. 5

Monthly Status Report 5

MILLIMETER WAVE RECEIVER STUDY

Contract period covered
1 August 1983 through 31 August 1983
P.O. No. X1173

A-3517
Submitted to
ITT
Electro-Optical Products Division
7635 Plantation Road
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Physical Sciences Division
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Georgia Tech Research Institute
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Atlanta, Georgia 30332

September 13, 1983

I. GENERAL

This work has continued to receive heavy emphasis during August. The major tasks treated were:

1) Definition of a mixer approach having sufficient flexibility a) to be incorporated into a variety of receiver architectures, b) to be fabricated by means either available or anticipated at Georgia Tech and ITT and c) to be capable of providing performance approaching the state of the art under the constraints of a) and b).

2) Definition of a planar mixer diode geometry compatible with the mixer design.

3) Preparation for review meeting with ITT personnel on 6 September, 1983.

4) Preparation of a tentative proposal for continuation of Georgia Tech efforts during the 1984-1985 time frame.

II. Receiver Architecture

Tasks 1 and 2 and the proposed work referred to in task 4 of the previous section (I) were derived from several considerations, some of which are potentially in conflict.

Consider first a totally integrated monolithic receiver containing all the desired semiconductors, filters and coupling elements. The most desirable approach, if one looks a few years into the future, is probably to employ several low noise FET amplifier stages followed by a FET mixer and I. F. amplifier with appropriate waveguide or planar ports and filters for the RF and LO and possibly a dielectric resonator stabilized FET local oscillator. (It is not clear that such an oscillator would be sufficiently immune to load pulling to omit ferrite devices).

Even if one assumes the approach to be viable, it requires processing of FET geometries having .25 micron gate length. Although we are presently working at Georgia Tech on such geometries, neither we nor ITT are in a position to utilize this technology in a routine manner at present.

Consider second the proposition that the overall receiver geometry is more dependent upon system interface than any other single factor. And, the system interface is not presently defined.

Consider third our joint goal to establish a technology that can be extended, in the future, to higher frequencies without "starting over".

Consider fourth, the goal to demonstrate an operating monolithic mixer having moderate performance in 1984.

The conflict between ideal and practical approaches evident in these four considerations might easily lead to a paralysis of our effort. Therefore, we have proposed the following relatively pragmatic approach. We believe it to be reasonable and appropriate if practicality is taken into account.

1) For the present, the GaAs area required should be kept within approximately 0.25 x 0.50 inch dimensions. A dimension narrower than 0.25 inch is highly desirable for one of the sides.

2) As many as possible of the device and circuit elements having major effects on either bandwidth or noise figure should be included in situ on the substrate.

3) In all likelihood system requirements in the near future will require a waveguide interface for the RF port. Therefore, an orthogonal mode approach, using the waveguide properties to eliminate at least one filter and its accompanying losses, is probably optimum. A large number of successful mixers of this type has been demonstrated between 20 and 200 GHz.

4) Suspended substrate stripline appears to be an optimum medium for the following reasons:

- o Losses are lower than microstrip.
- o Line widths are maintained at reasonable minimum widths in filter networks.
- o The impedance is higher for a given circuit function.
- o The suspended approach is shielded and reasonable approaches exist to reduce cost of the fixtures (shield).
- o The suspended substrate is compatible with transitions to other media such as microstrip so that, for instance, a microstrip LO can be incorporated at a later time.
- o The suspended substrate can utlimately incorporate other mixed media such as slotline, coplanar line and finline. Thus, the suspended approach has great versatility for transitioning to other system components.

Two uniquely simplified fixture designs for suspended substrate mounting have been devised. Assuming ITT approval, a portion of the continuing effort will be expended to develop one or both of these designs.

III. <u>Work Planned Prior to December 1983</u>

The following Design/Fabrication tasks are contemplated during this time period if mutually agreed.

- o Fabricate and test low pass filter (scaled and/or actual size).
- o Adjust filter design.
- o Fabricate and test hybrid integrated version of 30 GHz mixer using either GaAs or another dielectric having appropriate dielectric constant (ε_r = 12.9) for substrate.
- o Optimize fixture height and dimensions for diode match.
- o Choose between fundamental and subharmonic mixer designs (or develop both approaches for a later decision).

IV. Technical Discussions with ITT September 6, 1983.

This report, for the formal reporting period 1 August - 31 August has been prepared subsequent to our visit with ITT on Sept. 6, 1983. Some of the discussion herein take into account questions and other factors considered during that visit. Several questions raised during that visit have not been resolved. C. T. Rucker will visit ITT on either September 15th or 22nd to attempt to resolve remaining issues.

Del No. 6

Monthly Status Report 6

MILLIMETER WAVE RECEIVER STUDY

Contract period covered
1 September 1983 through 30 September 1983
P.O. Box X1173

A-3517
Submitted to
ITT
Electro-Optical Products Division
7635 Plantation Road
Roanoke, Virginia 24019

by

Physical Sciences Division
Electromagnetics Laboratory
Engineering Experiment Station
Georgia Institute of Technology
Atlanta, Georgia 30332

Contracting through
Georgia Tech Research Institute
Georgia Institute of Technology
Atlanta, Georgia 30332

October 10, 1983

I. GENERAL

A program review was held on September 6, 1983. During the review it became apparent that Georgia Tech was proceeding more directly than desired toward a specific mixer geometry and toward a specific mixer diode fabrication process.

Subsequent to the program review, C. T. Rucker again visited ITT to discuss project status and to restructure efforts in a manner more acceptable to ITT. Our present understanding of project emphasis is given below.

II. PROJECT EMPHASIS

1) ITT prefers a less pragmatic approach to the overall receiver design wherein detailed consideration is given to the various performance and fabrication tradeoffs inherent in the design of a millimeter wave monolithic receiver. For example, in the case of the local oscillator, the rationale for choice of either a fundamental or subharmonic approach, the means of stabilization or locking, the most viable fabrication medium, the type of active device etc. should be studied. The local oscillator is discussed here only as an example. Performance and complexity tradeoffs should accure from the study efforts.
2) An important portion of the project output is one or more suggested mask designs involving one or more particularly troublesome elements of the receiver.
3) Although the program completion date is unaltered, the

detailed schedule is altered in accordance with the new schedule attached. The final report delivery requirement is changed to January 31, 1984 in order to allow additional time for the technical effort.

4) It is anticipated that considerable tradeoff and design material for the local oscillator portion of the effort will be available from the separate local oscillator work already funded by ITT and that assistance with filter trade-offs will be available via Inder Bahl.

Georgia Tech considers the above restructuring of the effort to be acceptable. A significant portion of the trade-off activities can be effectively completed by the end of December. Completion of some of the trade-off activities will be completed during additional work contemplated for CY 1984. Mask design information related to critical items also appears to be a reasonable goal.

III. <u>RECEIVER APPROACH</u>

The basic assumptions with which we are proceeding are:

1) The substrate material is Gallium Arsenide,

2) the construction is monolithic, but hybrid development steps are appropriate,

3) for the long term all receiver semiconductor devices should be fabricable on a common substrate.

4) Receiver devices may include:

FETs - I.F. Amp.

FETs - R. F. Amp./Mixer (long term - except subharmonic)

GUNNS - Local oscillator

FETs - L.O. (long term)

MIXER - Diodes

5) Passive Receiver elements may include:

TRANSITIONS, FILTERS, COUPLERS, SPECIAL LINE GEOMETRIES, AIR BRIDGES, CAPACITORS AND INDUCTORS.

For the overall receiver approach, several considerations are common to all the receiver elements. These are:

1) Substrate area required

2) Physical compatibility of each element with remaining elements

3) Substrate losses

4) Semiconductor process compatibility among elements

Given these general considerations, the various receiver elements will be assessed in terms of rf performance, fabrication limitations and potential for realistic applications. For example, the mixer portion of the receiver will be studied as outlined below.

Mixer

For each mixer type considered, the constituent parts such medium; the device geometry; the device process etc. will be studied.

Potential mixer types include:

- o Single-ended Fundamental (single & dual diode)
- o Double-ended Fundamental
- o Balanced Fundamental
- o Balanced Crossbar-type
- o Double balanced
- o Image Reject
- o Image Enhanced
- o Balanced Subharmonic and
- o Harmonic

A similar overall view of the local oscillator, I.F. and RF port considerations will also be attempted.

IV. PROJECT FINANCIAL STATUS

Beginning Balance August, 1983	\$53,410
August Expenditures	\$13,498
Beginning Balance September, 1983	\$39,912
September Expenditures	\$13,610*
Beginning Balance October, 1983 (est.)	\$26,302
Estimated Permissible Expenditure rate through Dec. 1983	\$8500/mo.
Estimated Level of Effort	1 man month/month

*Includes = \$1700 M&S charge incurred in anticipation of hardware development during 1983.

V. <u>PLANS FOR OCTOBER</u>

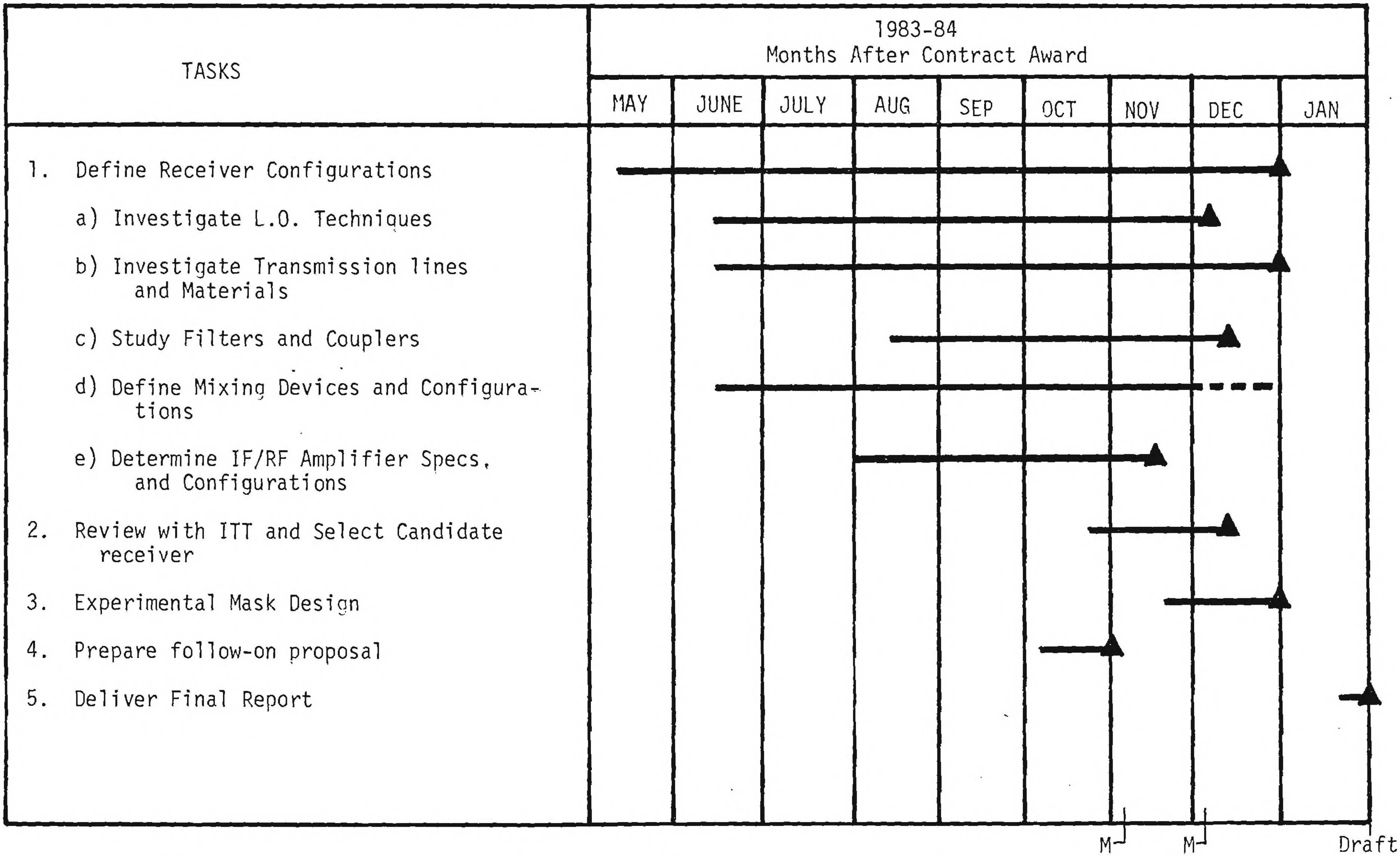
Proceed with study as described.

VI. <u>PROJECT SCHEDULE</u>

The attached project schedule has been revised in accordance with discussions with ITT.

TASK SCHEDULE

ITT - P.O. No. X1173



Revised 10/1/83

M - Meet with ITT

Del No. 7

Monthly Status Report 7

MILLIMETER WAVE RECEIVER STUDY

Contract period covered
1 October 1983 through 31 October 1983
P.O. No. X1173

A-3517
Submitted to
ITT
Electro-Optical Products Division
7635 Plantation Road
Roanoke, Virginia 24019

by

Physical Sciences Division
Electromagnetics Laboratory
Engineering Experiment Station
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Contracting through
Georgia Tech Research Institute
Georgia Institute of Technology
Atlanta, Georgia 30332

November 10, 1983

I. Introduction

This month, considerable progress has been made with respect to mixer/circuit studies and to local oscillator considerations. Conclusions noted appear reasonable based on current progress. Additional work is intended; therefore, additional conclusions or changes can be expected.

II. Local Oscillator

2.1 General Design Approach

The primary candidate devices for integrated local oscillator sources are the FET and Gunn devices. The FET is routinely fabricated on high resistivity substrates insuring its long term compatibility with mixer diodes and IF technology. Planar Gunn devices have been shown feasible as well, but these devices do not compare well to vertical geometries in terms of either efficiency or phase noise characteristics.

Fundamental frequency (30-35 GHz) FET local oscillators will likely require gate lengths significantly shorter than the half micron standard process devices presently being developed by ITT. Therefore, it is reasonable to consider either a subharmonically pumped approach or a fundamentally pumped approach using a half-frequency FET oscillator with doubler. This approach allows for use of half micron or longer devices. If necessary, some initial developmental efforts could employ vertical Gunn device structures integrated in a hybrid fashion. In this instance either the half-frequency or fundamental approach could be tolerated.

With either device, certain critical common problems exist. These include temperature instability, low Q and reproducibility. The first two result from the variation of device parameters with temperature and the relatively high noise associated with the low Q oscillator. Questions of yield are more critical in the local oscillator than in any other portion of the receiver. Filter losses, poor mixer diode quality, line width variations, material thickness variation and other factors could all contribute to nominal and gradual performance degradation. But, in the case of the local oscillator, a degree of adjustment must be provided simply to insure that the required frequency can be set.

It is not likely that any integrated monolithic L.O. approach will prove acceptable unless stabilized by some high Q and temperature stable reference means. Possible means include:

1) Phase or frequency locking by loops or injection from a quartz crystal derived stable reference,
2) Cavity stabilization by reaction or transmission techniques using either high Q metal cavities or dielectric resonators or
3) automatic frequency control (stabilization) by virtue of varactor tuning and a high Q discriminator network.

Each of these approaches offers its own set of advantages and disadvantages. Consider first oscillators that employ electronic tuning and a stable reference to obtain 1) midband frequency setting, 2) FM noise reduction and 3) temperature stabilization.

2.2 Electronically Tunable Oscillators

A) FET Doubling Oscillator

Using 1 micron NEC 69400 FETs, Winch[1] has reported a FET doubling oscillator tunable over the 19.5 to 30.6 GHz range. His schematic and performance are given in Figure 1.

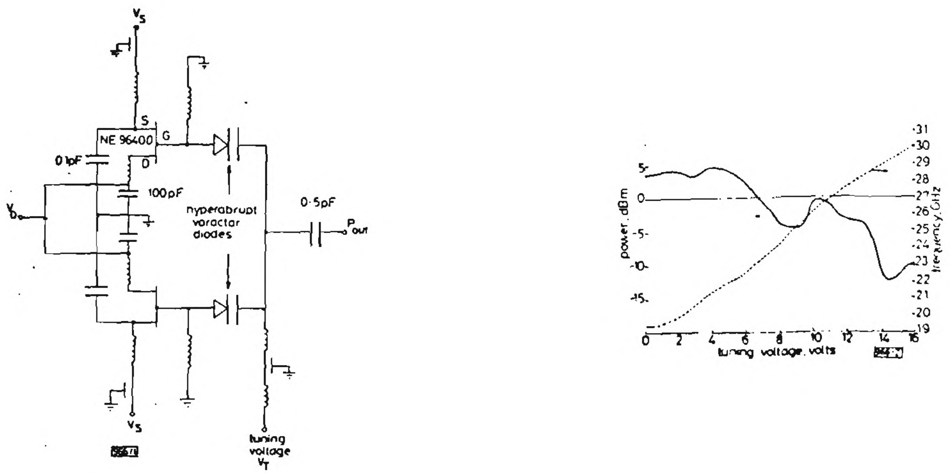

Figure 1. FET Doubling Oscillator (Winch[1])

The broad tuning range of this type of oscillator is a distinct advantage in that the same nominal design could prove useful for more than one application. Another advantage of such a broad tuning range is the ability to compensate readily for temperature frequency variations. Finally, by use of the doubling approach, FET requirements are far less stringent. A disadvantage is the requirement for the hyperabrupt varactors (see next paragraph).

This disadvantage is, however, no worse than that incurred with any half frequency source followed by a doubler. Another disadvantage is the power absorption in the varactor(s) when large tuning ranges are sought. The power roll-off problem must be solved regardless of the approach.

B) Monolithic Doubler

Chu et al[2] have recently reported a monolithic doubler employing planar Schottky barrier varactors. Their circuit provided 20-25% doubling efficiency with power output of about 20 mw at frequencies up to 27.4 GHz. Their circuit (Figure 2) has dimensions of 4 x 8 mm (.160 x .320 inch). No particular effort was made to fold or reduce the size of any of the circuit elements. With higher frequency of 35 GHz and other efforts to reduce size, it should be possible to fabricate the circuit in a region approximately 2.5 by 5 mm (.100 x .200 inch). The bandwidth achieved with this doubler indicates its use with an appropriate tunable oscillator.

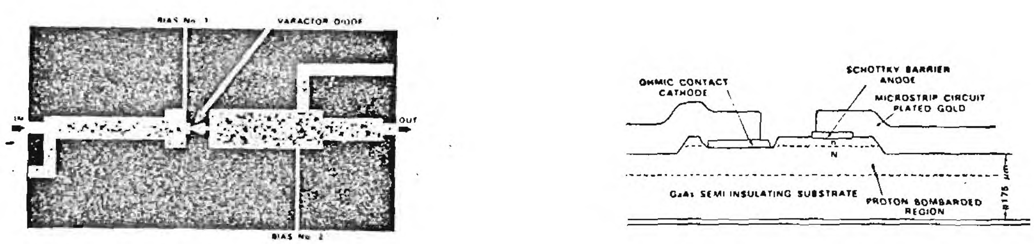

Fig. 2 FET Doubler (Chu et al[2]), Circuit and Planar Schottky Varactor

C) Rauscher[3] has reported both FET doublers and FET doubling oscillators using Avantek M106 chips. He predicted and achieved conversion gain of -1dB for an input level of +10dBm at 30 GHz in a FET doubler. His doubling oscillator provided 8.8 dBm of output at 29 GHz with 8.9 percent efficiency.

These three examples (A, B, C) appear to indicate that FET or Varactor doublers or FET doubling oscillators can provide the necessary tunability and power output levels to drive fundamental frequency mixers in Ka-band. Discussions with R. Trew indicate that a half frequency oscillator with doubler may presently be the most desirable approach with respect to noise performance.

D) Tunable Gunn Oscillator

Rubin[4] has reported an example of a typical varactor tuned Gunn oscillator for 35 GHz (Figure 3). His oscillator provided approximately 5 milliwatts of output power over a 1 GHz range as noted in Figure 3.

The oscillator is fabricated on a low dielectric constant substrate ($\epsilon = 2.3$) and has dimensions of about .90 x .75 inch including an extensive ground plane. Similar results could be achieved, using different grounding techniques, on a substrate

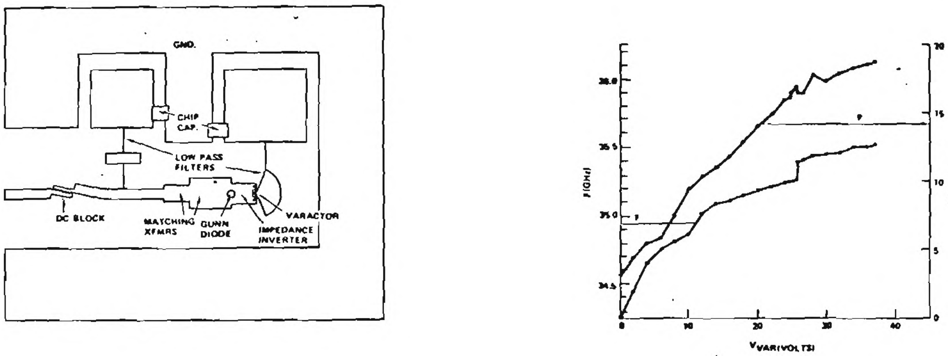

Figure 3. Varactor Tuned Gunn Oscillator (Rubin[4])

about .75 x .50 inch. With gallium arsenide substrates such a design would reduce to approximately .3 x .2 inch, still a rather large size for integration with the remaining L.O. and receiver elements.

Given the foregoing examples (many others are available) it is reasonable to take the view that a tunable local oscillator can be developed. If so, several possible means of stabilization and/or FM noise reduction exist.

2.3 <u>Frequency Control/Locking</u>

A) General Comment

It is usually necessary, where phase locked arrangements are used, to specify a range of sideband frequencies over which stabilization is required. The complexity of the stable reference and feedback loop are greatly affected by these specific requirements. In the next few paragraphs no attempt is made to consider any specific example. Instead, several nominal approaches that may prove useful are discussed.

B) Subharmonic Injection Locking

It has been shown that signals injected at a subharmonic of the output frequency can give results similar to that obtained with fundamental frequency locking. Unfortunately, locking gain and bandwidth are both quite low for this scheme. Its use is not recommended.

C) Parametric Injection Locking (PIL)

Okamoto[5] has demonstrated a reasonable circuit for parametric injection locking using an IMPATT operated at 75 GHz. He achieved noise reduction of about 25 db (as noted on a spectrum analyzer) and frequency stability as good as $10^{-8}/^{o}C$. Locking range of 700 MHz (1%) was obtained with injected power of 23 dBm. His circuit is shown schematically in figure 4.

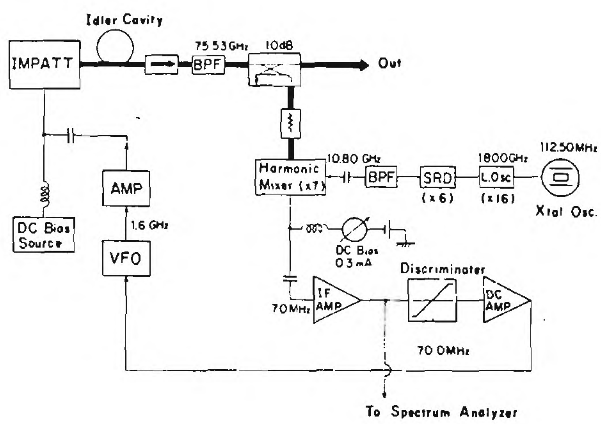

Figure 4. Parametric Injection Locked Oscillator Schematic (OKAMOTO[5])

Parametric injection locking offers the advantage that a low- frequency stable-oscillator can be employed for locking. A significant disadvantage is that a high Q idler resonator, operating near the output frequency, is also required to achieve his results. Furthermore, the stability and/or noise reduction depend on both the cavity and the stable low frequency signal. This approach is an unlikely choice at present.

D) Modified Pound[6] Circuit

Chan and Cole[7] have reported a stabilized Gunn oscillator in X-band using a circuit similar to that presented by Pound[6] in 1946. At X-band, they achieved an average short term stability of 2×10^{-7} and temperature stability of $9 \times 10^{-7}/^{\circ}C$ between -15 and $+30^{\circ}C$. The nominal oscillator and electronic circuits are shown in figure 5 for reference purposes only. We do not suggest that such a complex approach should be attempted at present. However, for sophisticated requirements, this level of complexity may be required. For less sophisticated requirements a relatively simple circuit such as that discussed next may suffice.

E) Discriminator Stabilized Oscillator

Glance and Snell[8] have described a very simple approach to a microstrip discriminator for use in C-band (figure 6). The circuit is just the usual S-curve discriminator achieved by use of a microstrip resonator having unequal rectangular dimensions. When coupled with a varactor tunable Gunn oscillator, this discriminator provided a stabilization factor of 1000 up to 1 KHz where the I.C. amplifier gain response began to decrease. The

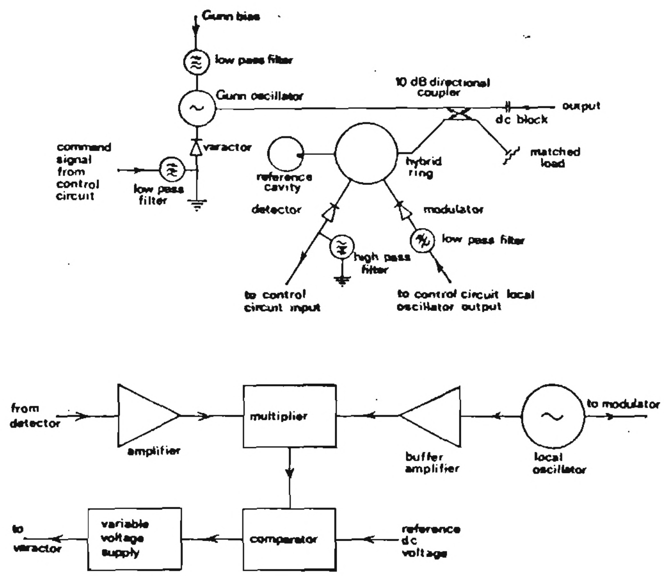

Figure 5. Modified Pound Circuit (Chan[7])

stabilized oscillator, including the I.C. amplifier, is shown in figure 7 along with the response characteristics of the discriminator.

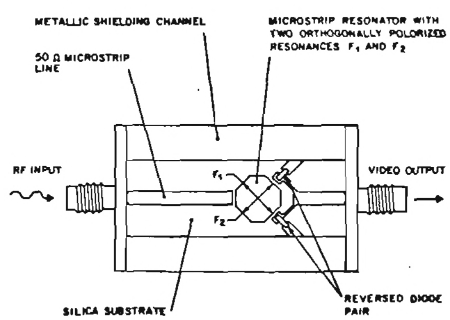

Figure 6. S-Curve Microstrip Discriminator

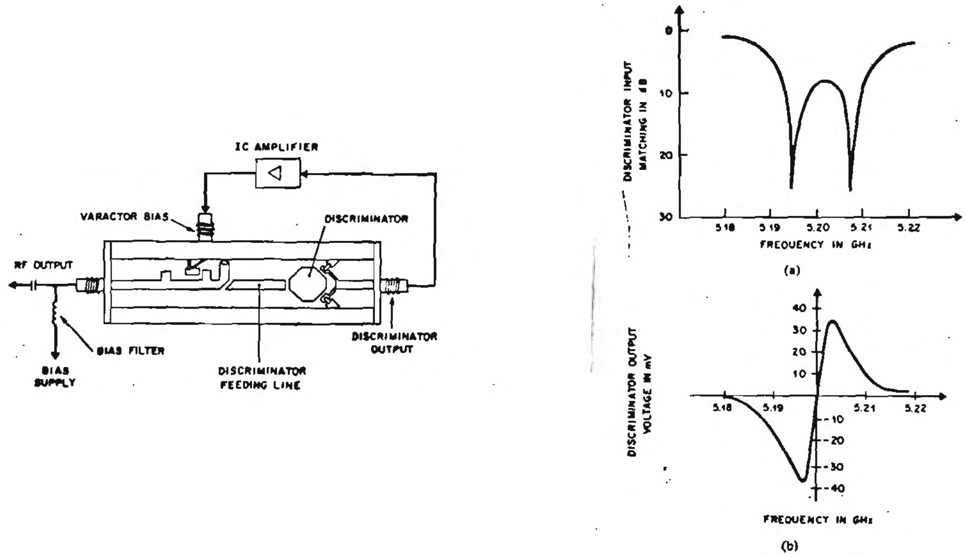

Figure 7. Stabilized Oscillator and Discriminator Characteristics.

In this particular example, the maximum voltage output of the I.C. was limited to a saturated value which would never cause the oscillator to tune outside the capture limits of the S-curve and saturated amplifier combination. Therefore no search mode was required within the closed loop. For this simple approach to be useful, the total frequency excursion of the basic low Q oscillator, due to manufacturing tolerances, temperature or any other effect, must not exceed the range of frequencies over which the amplifier's output is saturated. With reduced S-curve slope and more complex feedback electronics, even this range could be extended.

Given the simplicity of this basic approach, it is interesting to consider use of a discriminator employing an assymmetric rectangular dielectric resonator to achieve either a K_u or K_a-band discriminator. This discriminator could be coupled readily with any of the electronically tuned oscillators already discussed. An ideal combination might ultimately prove to be the discriminator coupled with the FET doubling oscillator suggested in paragraph 2.2.A or with a varactor tuned half frequency L.O. near 15 GHz.

Fixed frequency dielectric resonator stabilized oscillators will be discussed in detail in the next monthly status letter. It is interesting to note here, however, that the discriminator scheme might allow one to first prestabilize the oscillator using a fixed tuned dielectric resonator and then to further stabilize the oscillator in selected IF bands by use of the discriminator and a band limited feedback amplifier. It is our plan to look further into this scheme as well as fixed tuned dielectric cavity stabilized oscillators during the next month.

2.4 Tentative Conclusions/Local/Oscillator

1. Half micron FET geometries allow use of the FET as either a half frequency subharmonic pump or a half frequency voltage tuned doubling oscillator. A separate doubler is also feasible.
2. Addition of either active stabilizing circuits (locked loop, discriminator) or passive elements leads to overall circuit dimensions on GaAs that range up to 0.5 inch. Therefore, initial design and or fabrication work on the local oscillator will probably need to be done

using a separate GaAs substrate. More compact designs would then be developed.

3. Useful information related to both passive and active stabilization can be acquired using vertical geometry Gunn device chips mounted adjacent to the remainder of the L.O. system (e.g. planar varactors, planar detectors, discriminator circuit, dielectric resonators).

4. These tentative conclusions, based on the literature and our assessment of the present technology are compatible with R. Trew's current thinking with respect to the L.O. approach.

III. Mixer/Circuit Study

3.1 Work Scope

During this period the mixer circuit portion of this study has consisted of : 1) reviewing articles and work performed by other researchers; 2) evaluating suitable transmission media; 3) performing circuit design tradeoffs of some of the required mixer circuits; and 4) determining the types of circuits required for each mixer type.

3.2 Literature Investigation

During the review of articles to date it has been found that most of the monolithic receiver work that has been done using very simple circuits such as the crossbar mixer which has two diodes, a low pass filter and a waveguide to transmission line (suspended substrate stripline) transition.[9,10] Another type of monolithic circuit has also been developed at 31 GHz using an integrated coupler.[11] This mixer requires a relatively large GaAs chip size because of the presence of coupler which combines the LO and RF frequencies. Diagrams of single ended and crossbar types of mixer are given in Figures 8 and 9. Conversion losses of between 4.5 and 12.0 dB have been achieved over broad bands at 30 and 60 GHz with the crossbar versions using suspended substrate stripline circuits on GaAs substrates.

3.3 Media and Filters

An in depth study of transmission line media has been performed. Details of this study will be included in an Appendix in the final report. The findings are that the optimum

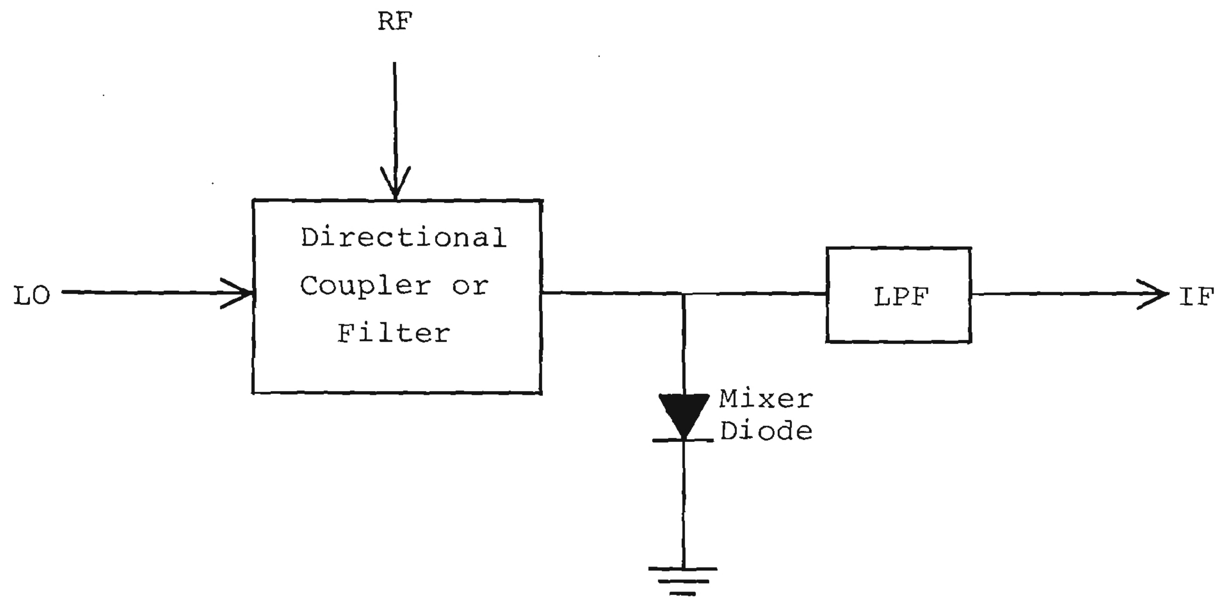

Figure 8. Single Ended Mixer Diagram

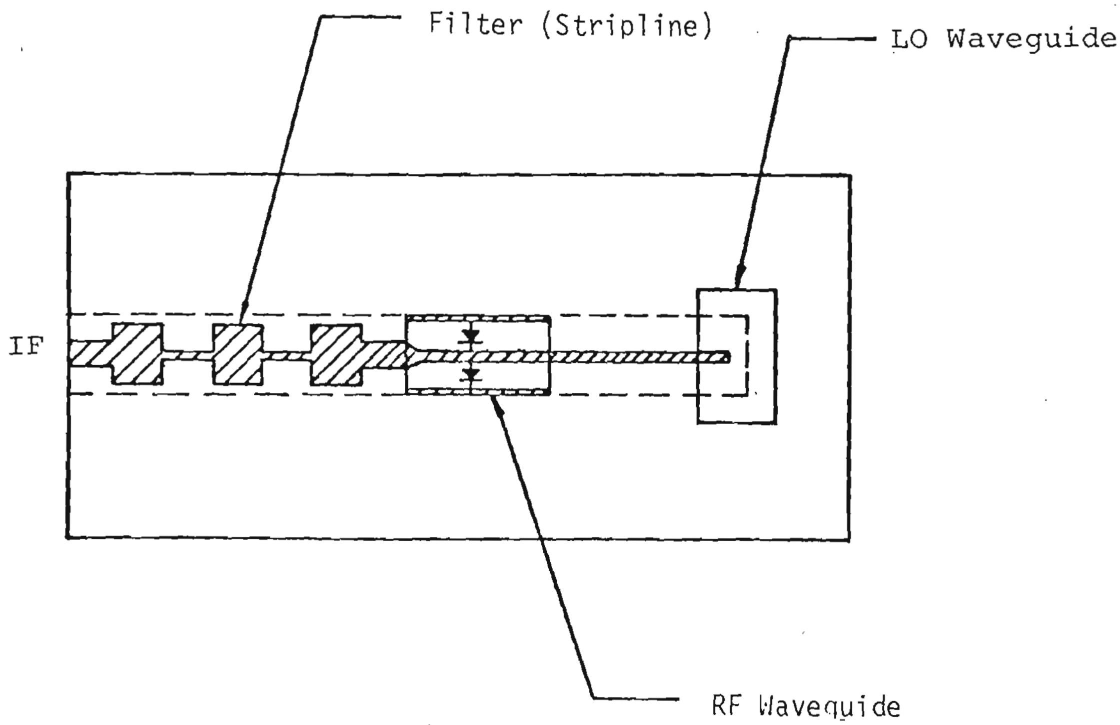

Figure 9. Crossbar Mixer Diagram

transmission line medium suitable for use in developing K_a -Band monolithic mixer circuits is either microstrip or suspended substrate stripline. These can sometimes be conveniently combined with waveguide to achieve some of the balanced configurations. Probes can also be used as transitions to waveguide. The final choice of either of these particular media depends highly on the mixer type, circuits, and coupling to RF, LO, and IF ports. For instance, a mixer that requires a directional coupler, a rat race hybrid, or a directional filter would need to be made in microstrip. In this case, a suspended substrate approach would lead to very large area substrates by comparison to microstrip.

A low pass filter is required in almost every mixer type to separate the RF and LO energy (both near 30 GHz) from the IF (at about 7 GHz). Tradeoffs can be made in: 1) choice of cutoff frequency; 2) number of elements; 3) impedance values of semi-lumped elements ; and 4) choice of media or medium.

Assume, for example, a low pass filter circuit fabricated on a 0.010" thick GaAs substrate. The two preferred media, suspended substrate stripline, and microstrip are shown in Figure 10. Plots of characteristic impedance and phase velocity for these two transmission lines are given in Figures 11 and 12. It can be seen from these graphs that higher impedances are more difficult to achieve on microstrip than on suspended substrate stripline. This lowers the reasonable values of achievable high impedances for microstrip which are needed to make series inductors for the low pass filter; a lower impedance value means that longer lengths of

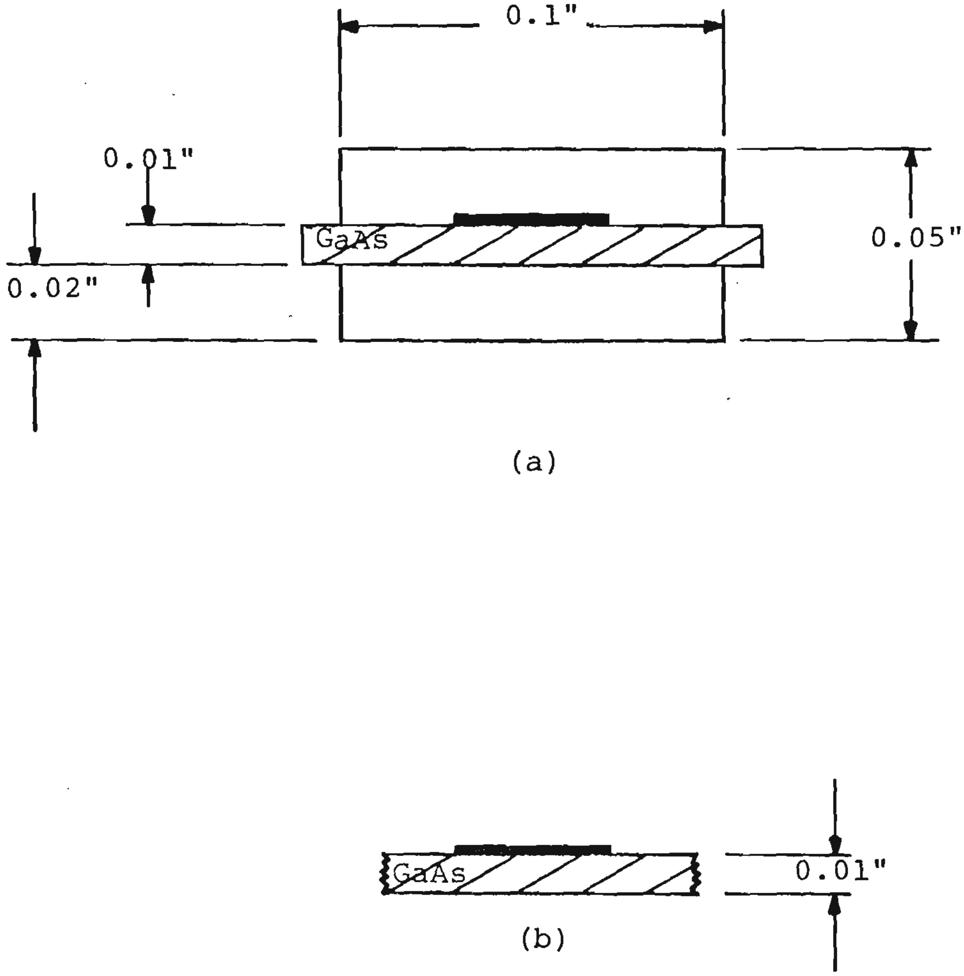

Figure 10. (a) Suspended Substrate Stripline and
(b) Microstrip Cross Sections.

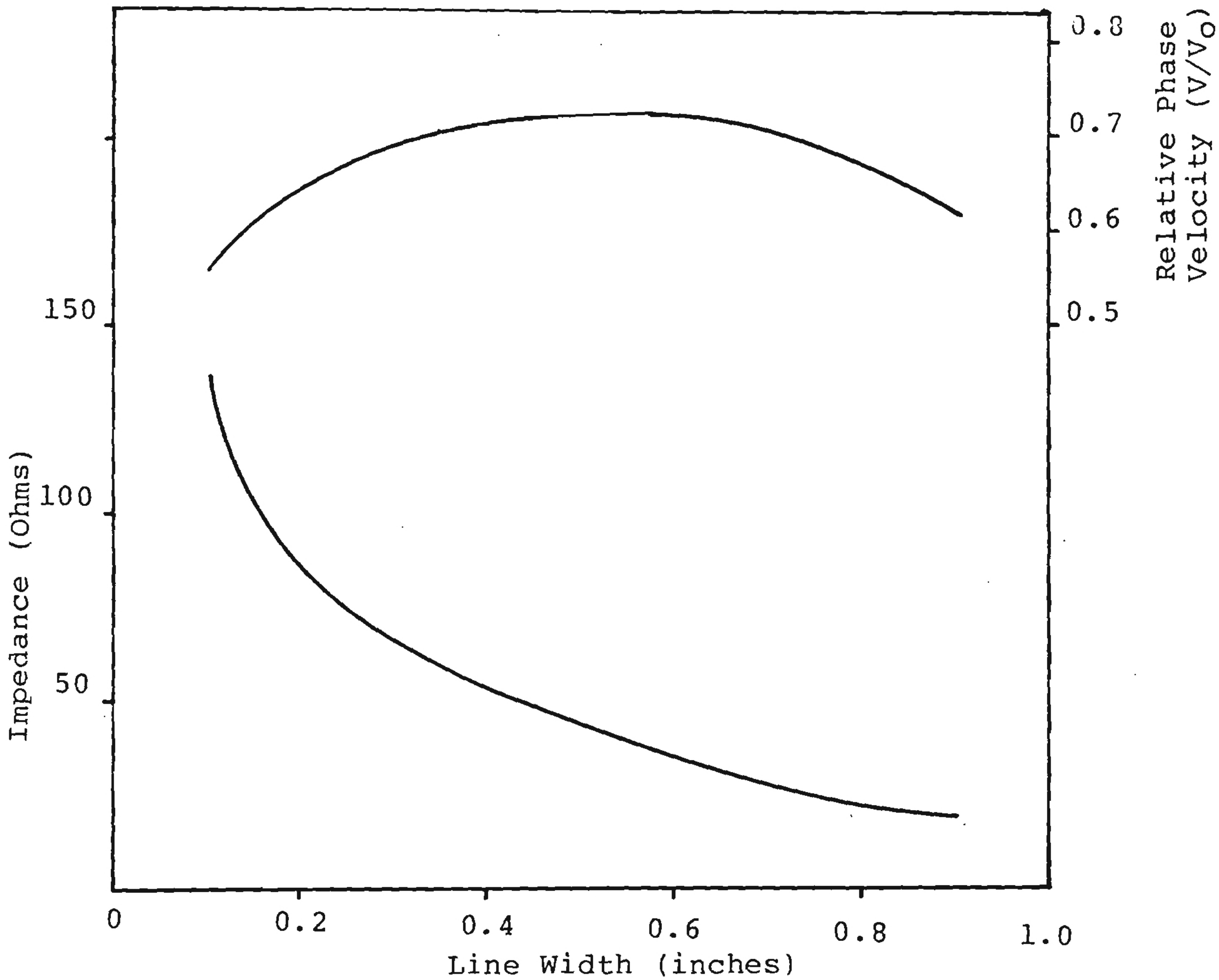

Figure 12. Characteristic Impedance and Phase Velocity Versus Line width for Suspended Substrate Stripline with a 0.01" GaAs Substrate

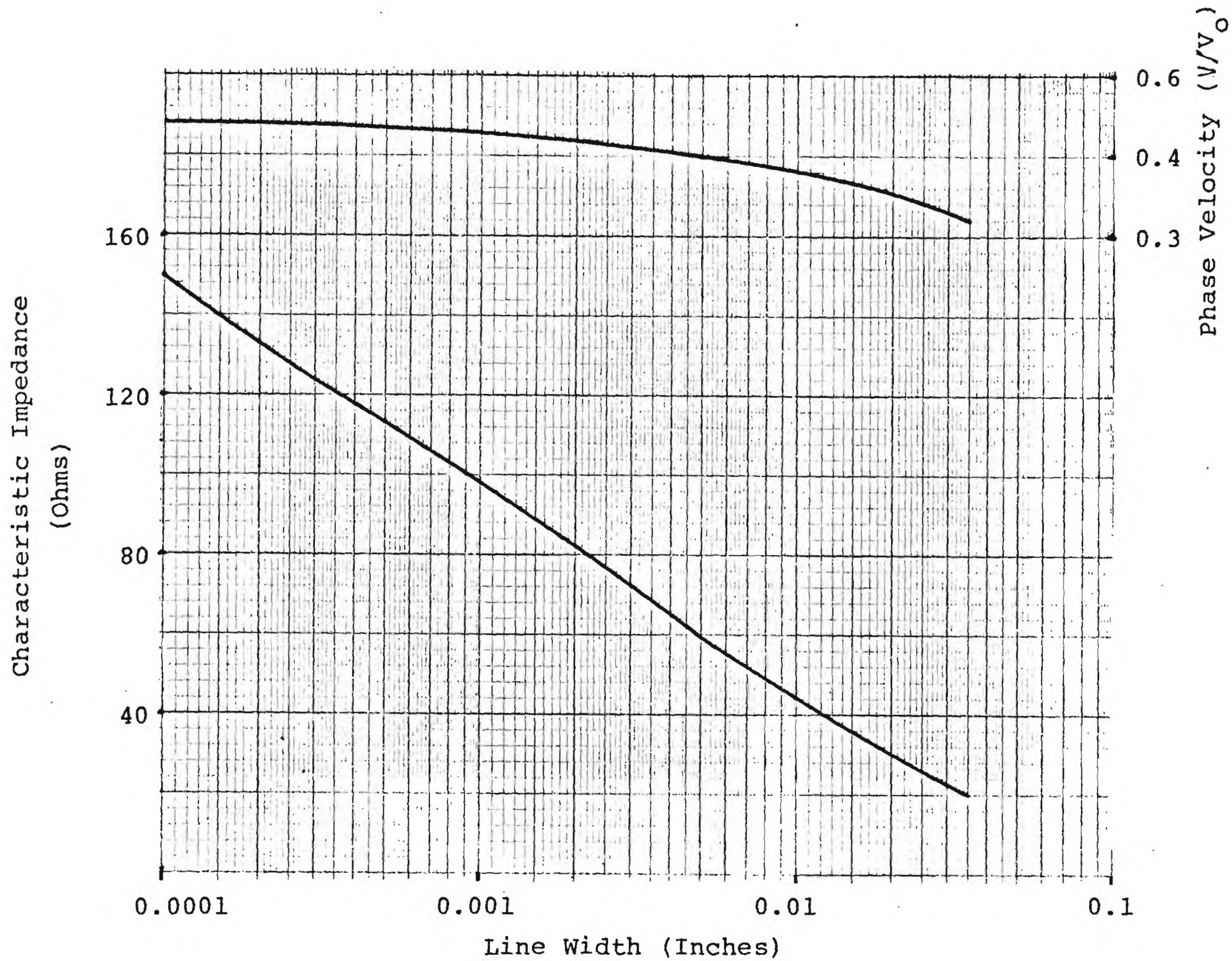

Figure 12. Characteristic Impedance and Phase Velocity for 0.01" GaAs Microstrip.

line are needed for a given inductance. An upper limit in length is reached when the length approaches a quarter wavelength at the highest operating frequency. The highest reasonable impedance in microstrip would be about 95 ohms or a 0.001 inch line width. However, 150 ohm line is achievable with suspended substrate stripline. Some filter responses showing the tradeoffs in the filter response for 5 and 7 element filters using either 95 or 150 ohm inductors are given in Figure 13. Overall filter length is also provided for both microstrip and stripline versions of these filters. The size of the filter helps determine overall receiver chip size. However, placement of the filter can be critical for optimum mixer operation and this effects chip size as well.

3.4 Mixer Types

Determination of the types of circuits needed for each mixer is required to help estimate chip size and mixer performance. Some of the mixer types and their required circuits are listed in Table I. This table is not complete but should be completed prior to the next reporting period. Additional circuit tradeoffs and preliminary mixer circuit designs will also be presented in the next report.

3.5 Progress With FET Amplifiers/Mixers

Great progress is being made in the development of monolithic FET amplifers and mixers at frequencies rapidly approaching K_a-band. Some current examples were recently provided at the GaAs IC Symposium held in October.[12] Such devices are of great interest and should be considered for the future. FET amplifers and/or mixers are not considered herein.

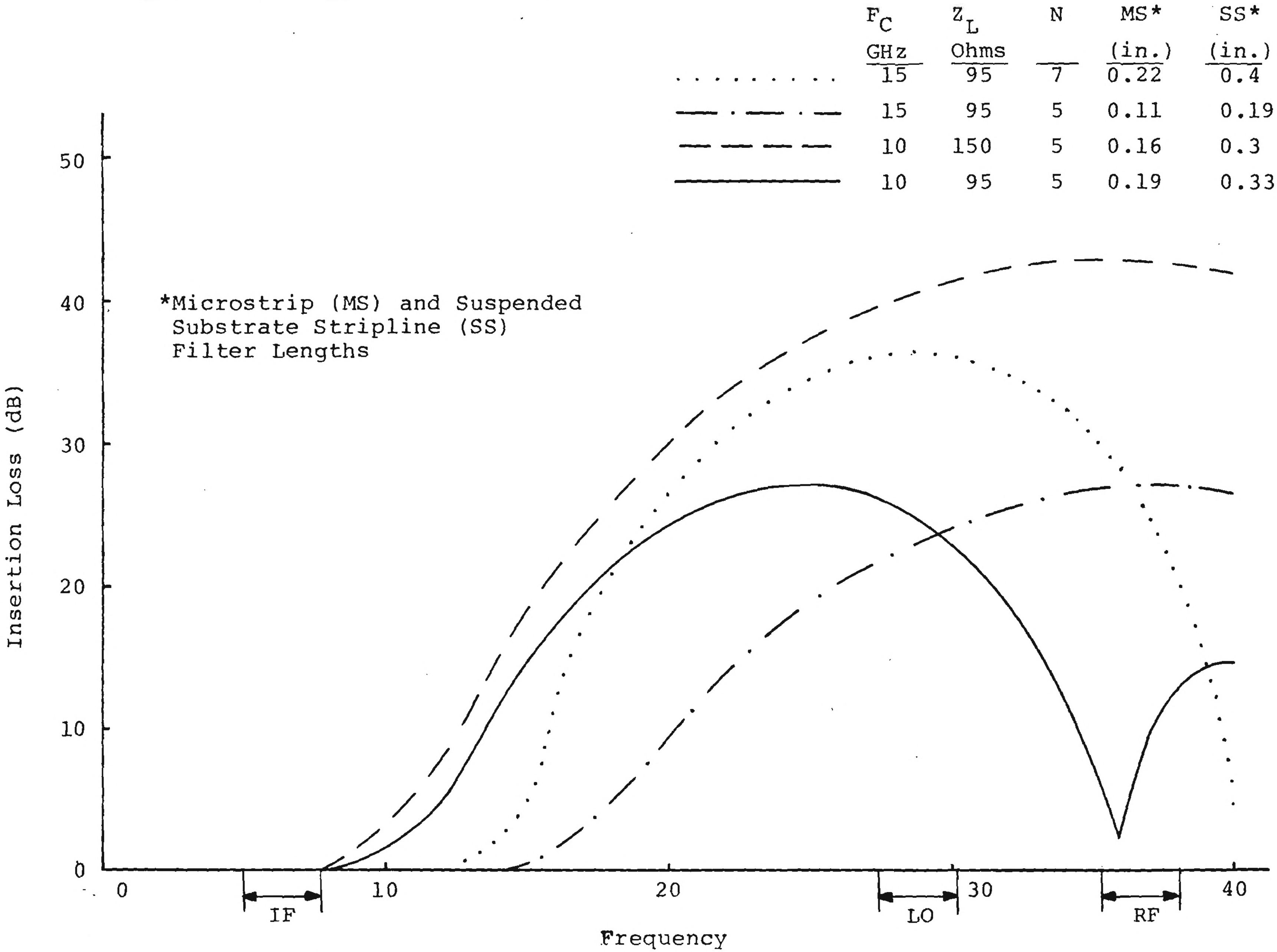

Figure 13. Filter Responses for Various Inductor Impedances (Z_L), Cutoff Frequencies (F_C), and Number of Elements (N).

TABLE I

Mixer Type	Circuits Required
Single Ended Fundamental	LPF/Transition/Couplers Ring Resonator direct filters
Double Ended Fundamental	LPF/Transition or BPF or diplexer
Balanced	3dB coupler/Rat Race Hybrids LPF, transitions
Balanced Crossbar	LPFs/Transitions to wave-guide/H-Plane IF/LO Feed of two Diodes in RF waveguide
Image Reject (Low band)	Place HPF in front of RF port of any of these other mixers
Image(Sum Enhanced) (2 x LO - RF)	To be specified
Balanced Subharmonic	LO/IF diplexer, LPF, WG to stripline transition
Harmonic	LO/IF diplexer, LPF, WG to stripline transition

IV REFERENCES

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9) Chao, et al "Ka-band Monolithic GaAs Balanced Mixers," IEEE Trans. on MTT, Vol. 31, No. 1, Jan. 1983.

10) L. T. Yuan and J. A. Paul, "High Cutoff Frequency Beam Lead Diodes and Mixers," Interim Reports, ERADCOM Contract, DAAK20-81-C-0417, January and March 1983.

11) A Chu, et al "A 31 GHz Monolithic GaAs Mixer/Preamplifier circuit for Receiver Applications," IEEE Trans. Electron Devices, Vol. ED-28, No. 2, pp 149-154, Feb. 1981.

12) Technical Digest, GaAs IC Symposium, October 25-27, 1983, pages 3-6, 11-12, 13-15, 20-23.

V. Tasks Suggested for 1984

Suggestions for 1984 tasks were discussed with ITT on October 31, 1983. Three primary areas of work were discussed, They are:

- o Design, development and test of mixer diode fabricated on high resistivity substrate using ion implantation.

- o Design and development of critical components such as filters, hybrids (if used), transitions and treatment of IF amplifier-diode interface.

- o Integration of above elements to the extent feasible within funding level.

If possible within project scope an additional task appears to have potential at present. If an S-curve discriminator approach is ultimately developed for the L.O., initial design and optimization of the dielectric resonator/discriminator circuit can be carried out independent of other L.O. or receiver components and would constitute a significiant development step.

VI. PLANS FOR NEXT MONTH

- o Additional mixer circuits tradeoffs and a preliminary mixer design will be accomplished.
- o Fixed frequency dielectric resonator oscillators will be studied in detail and discriminator stabilization will be considered further.

VII. PROJECT FINANCIAL STATUS

Beginning Balance, October, 1983 (actual	$24,574
October Expenditures	6,340*
Beginning Balance, November, 1983	18,234

*Actual expenditures of approx. $9000 were reduced to $6340 by credit applied to correct erroneous charges debited in September.

A-3517

Final Technical Report
Project A–3517

MILLIMETER WAVE RECEIVER STUDY

By

Physical Sciences Division
Electromagnetics Laboratory
Engineering Experiment Station

Prepared for

INTERNATIONAL TELEPHONE & TELEGRAPH
ELECTRO–OPTICAL PRODUCTS DIVISION
7635 PLANTATION ROAD
ROANOKE, VIRGINIA 24019

Under

Purchase Order Number x1173

February 20, 1984

GEORGIA INSTITUTE OF TECHNOLOGY
A Unit of the University System of Georgia
Engineering Experiment Station
Atlanta, Georgia 30332

1984

FINAL

TECHNICAL REPORT

MILLIMETER WAVE RECEIVER STUDY

A-3517
P.O. No. X1173

Prepared for

ITT
Electro-Optical Products Division
7635 Plantation Road
Roanoke, Virginia 24019

by

Physical Sciences Division
Electromagnetics laboratory
Engineering Experiment Station
Georgia Institute of Technology
Atlanta, Georgia 3033

Contracting through

Georgia Tech Research Institute
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Atlanta, Georgia 30332

February 20, 1984

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

I. INTRODUCTION

The purpose of this work has been to examine various alternates of design approach, fabrication media, local oscillator types and stabilization means leading to a monolithic receiver approach for operation near 35 GHz. Included are discussions related to the overall receiver topology; mixer types, circuits and designs; local oscillator approaches and means of frequency stabilization.

II. MONOLITHIC RECEIVERS

2.1 Background

Considerable interest has been given to the development of millimeter wave monolithic receivers. However, very little has yet been accomplished in this area. The majority of the work to date has been in the development of mixers and IF amplifiers with some work incorporating monolithic antennas.

Monolithic receivers are being developed for direct broadcast satellite receivers as high as 12 GHz. Each of the major components has been successfully developed and tested.(2-1, 2-2) S. Hori et. al. have taken this development one step further by testing an integrated receiver over the 11.7 - 12.2 GHz band with an overall noise figure of 4.0dB. This receiver uses three monolithic chips.(2-1, 2-2, 2-3)

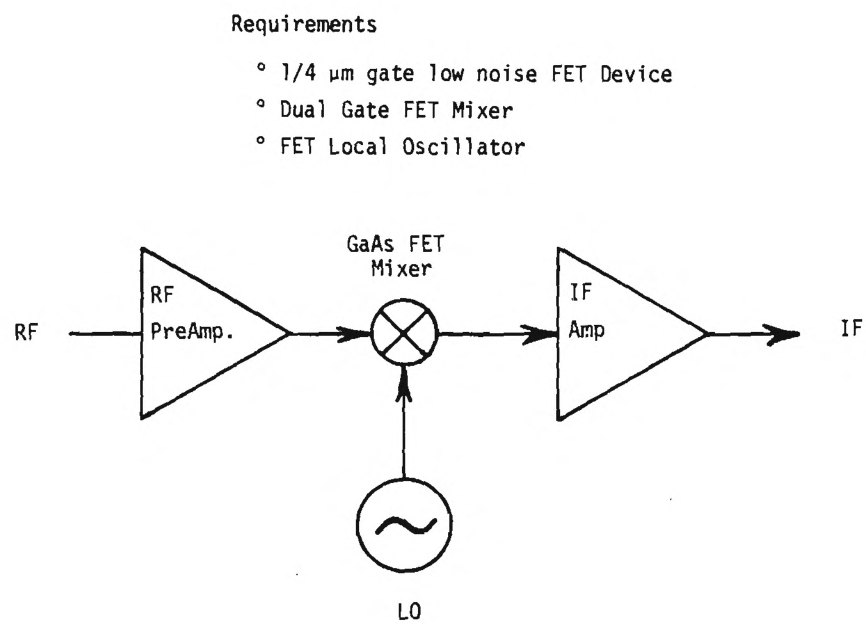


Figure 2-1. Block Diagram of Monolithic All GaAs FET Receiver.

A block diagram of these receivers, is shown in Figure 2-1. Each uses low noise GaAs FET preamplifiers, dielectric resonator stabilized FET local oscillators, FET IF amplifiers and dual gate GaAs FET mixers. Other passive devices that been developed on GaAs for use in these receivers include inductors, capacitors, couplers and resistors. These devices have been realized using microstrip as the transmission medium.

This technology is extendable to higher frequencies such as 35 GHz. However, the key component, (a low noise GaAs FET preamplifier at 35 GHz), requires a quarter micron gate length.

Because of this difficulty low noise monolithic receivers at 35 GHz and higher will likely utilize low noise diode mixers as the key receiving element for the near term. The crossbar mixer has been investigated by both L. Yuan (Hughes) and C. Chao (TRW). This mixer type has been chosen because of its simplicity (in terms of number of circuits) thus reducing overall size, performance, and bandwidth.[2-4, 2-5]

C. Chao's work has been at 35 GHz. A 4.5dB double sideband noise figure has been achieved using a single GaAs substrate in a suspended stripline configuration over a 30-32 GHz bandwidth. A 6 dB noise figure has been achieved over a 31-39 GHz frequency range. This includes a 1.5 dB IF noise figure. The overall circuit size was 0.05 x 0.43 inches. This includes the RF, LO, and IF mixer circuitry but not the IF or LO devices themselves. An RF waveguide input with an H-Plane IF/LO feed is the key to this balanced mixer's operation. The LO can be provided via a transmission line medium such as microstrip.

Other monolithic mixer/antenna work has been performed at higher frequencies. A silicon dielectric waveguide antenna has been integrated with a silicon mixer diode with low efficiency at 85 GHz by C. Yao et. al. Also a planar diode/probe has been developed for use in a cooled 110 GHz system by B. Clifton et. al. with a conversion loss of 4-5dB. Quasi-optical devices were used to couple the external LO to the diode probe.(2-6, 2-7)

2.2 Receiver Topologies

The simplest approach to superheterodyne receiver design uses the single ended mixer as shown in Figure 2-2. It requires

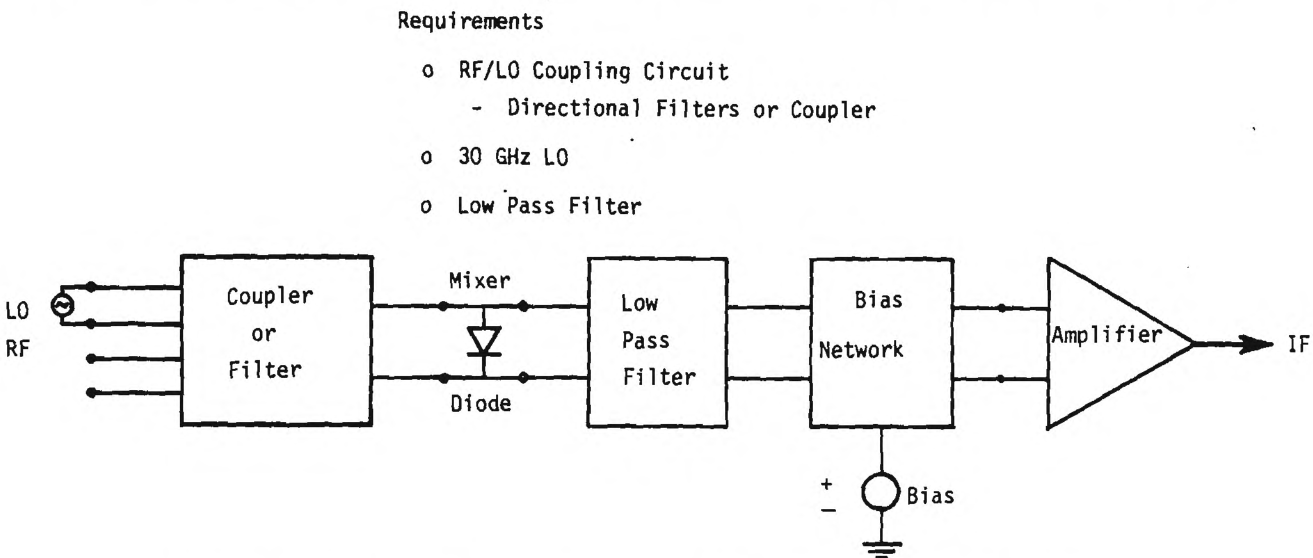


Figure 2-2. Single Ended Mixer Receiver Topology.

only one diode, an RF/LO coupling circuit such as a directional coupler or filter, a low pass filter, a fundamental (30 GHz) local oscillator, and possibly a bias circuit.

A balanced mixer receiver, such as that shown in Figure 2-3, requires two diodes and a hybrid circuit in addition to a 30 GHz LO and an RF/LO choke (low pass filter). The hybrid circuit can be realized using a Lange coupler, a rat race hybrid, a direct coupled hybrid, or a crossbar configuration (which uses an H-plane fed RF waveguide).

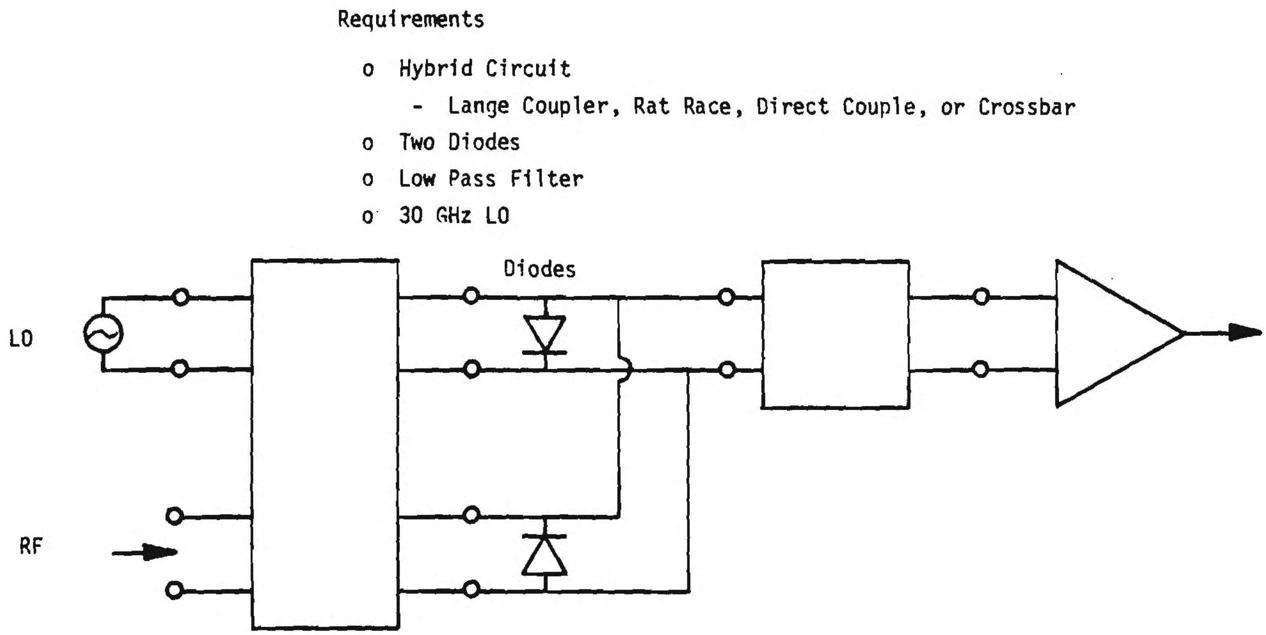


Figure 2-3. Balanced Mixer Receiver Topology.

The LO and RF are split by the hybrid circuit and fed to the two diodes 180 degrees out of phase. The Lange coupler and direct coupled hybrids require an extra length of line for proper phasing since they are 90 degree hybrids.

Receivers that use subharmonically pumped mixers, such as that shown in Figure 2-4, have several advantages over the other receiver types. It is a balanced mixer that will suppress local oscillator AM noise and requires two diodes in an antiparallel configuration. It's main feature is that the LO frequency is about one half the signal frequency. This eases the local oscillator requirements considerably. This is significant in that the necessary monolithic LO near 30 GHz has not yet been developed.

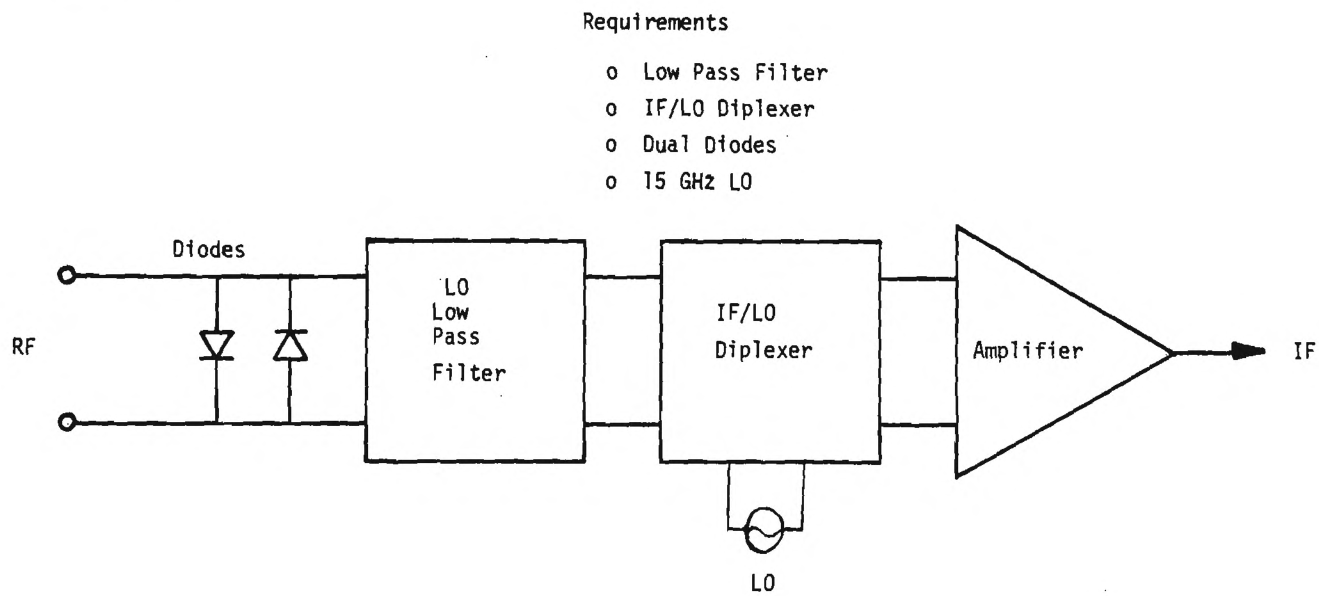


Figure 2-4. Subharmonic Receiver Topology.

The antiparallel diode configuration reduces conversion loss to values comparable to fundamental mixers because the fundamental mixing products are suppressed by the diode circuit itself. Excellent results have been obtained with these mixers as high as 220 GHz.(2-8, 2-9, 2-10)

Other mixer types such as image/sum enhanced mixers are not considered in this study because of their complexity, (large number of external circuits), in spite of their excellent performance (2.8 dB conversion loss at 9 GHz). Double balanced mixers also require sophisticated circuitry without any significant advantages over the other receiver types discussed earlier.[2-11, 2-12, 2-13, 2-14]

Harmonic mixers make simple receivers and use only one diode. They also employ a half frequency LO but exhibit a higher conversion loss. Harmonic mixers might be suitable if preceded by an RF preamp to lower the receiver noise figure. The circuitry required for the harmonic mixer is similar to the subharmonic mixer shown earlier but would likely require a dc bias as well.

III. MIXERS

3.1 Mixer Types

Many types of mixers have been developed for use in the millimeter wave frequency range. Those described next are: 1) single ended mixers; 2) balanced mixers; and 3) subharmonically pumped mixers. These are chosen primarily because of their high potential for use in monolithic form.

The most common and simplest type of mixer is the single ended type in which the LO and RF are injected via a single input port. This mixer type usually requires only one diode. A dc bias may or may not be required for optimum mixer operation depending on the diode's IF and RF characteristics (noise, conversion loss, and impedance mismatch) when pumped. A diagram of such a mixer and the circuits required was shown in Figure 2-2. For monolithic applications all of these circuits should be fabricated on GaAs in microstrip or suspended substrate stripline.

The RF and LO are injected into the mixer diode using either a directional coupler or a directional filter. RF and LO matching can be done using open or short circuited stubs and an RF choke (low pass filter). Proper attention must be given to the impedances seen by the diodes at multiples of the LO frequency to reduce conversion loss. This can be done using a computer program developed by A. Kerr along with a suitable diode circuit model. The diodes developed for use in these receivers should be carefully measured to determine their equivalent circuit parameters.[3-1, 3-2]

An IF matching circuit is usually required to reduce losses and match the IF energy to the input of the IF amplifier. This can be accomplished using either lumped or semi-lumped matching techniques, quarter wave transformers, or stub matching. The bias circuit usually consists of a series dc-blocking capacitor and a parallel inductor. This mixer type is the only one which typically requires this dc bias circuit. The effect of bias helps reduce LO requirements and assists in impedance matching the diode to the external mixer ports.

The balanced mixer configuration has been shown earlier in Figure 2-3. The types which use special hybrid circuits to achieve the balanced configuration are straight-forward in their application as noted previously.

The crossbar mixer types, however, have several interesting qualities. In particular, the balanced configuration is achieved using an H-plane IF/LO feed into an RF rectangular waveguide as shown in Figure 3-1. This arrangement puts the diodes in series as seen by the RF and in parallel as seen by the IF and LO. This series RF arrangement increases the impedance of the diode pair to help match the impedance of the RF energy. The parallel IF arrangement also lowers the impedance of the diodes as seen by the IF port thus lowering the IF mismatch losses. Extremely broadband, low noise receivers have been made using these types of mixers. No special LO/RF hybrid circuit is required and the RF is never transmitted on a transmission line medium which helps

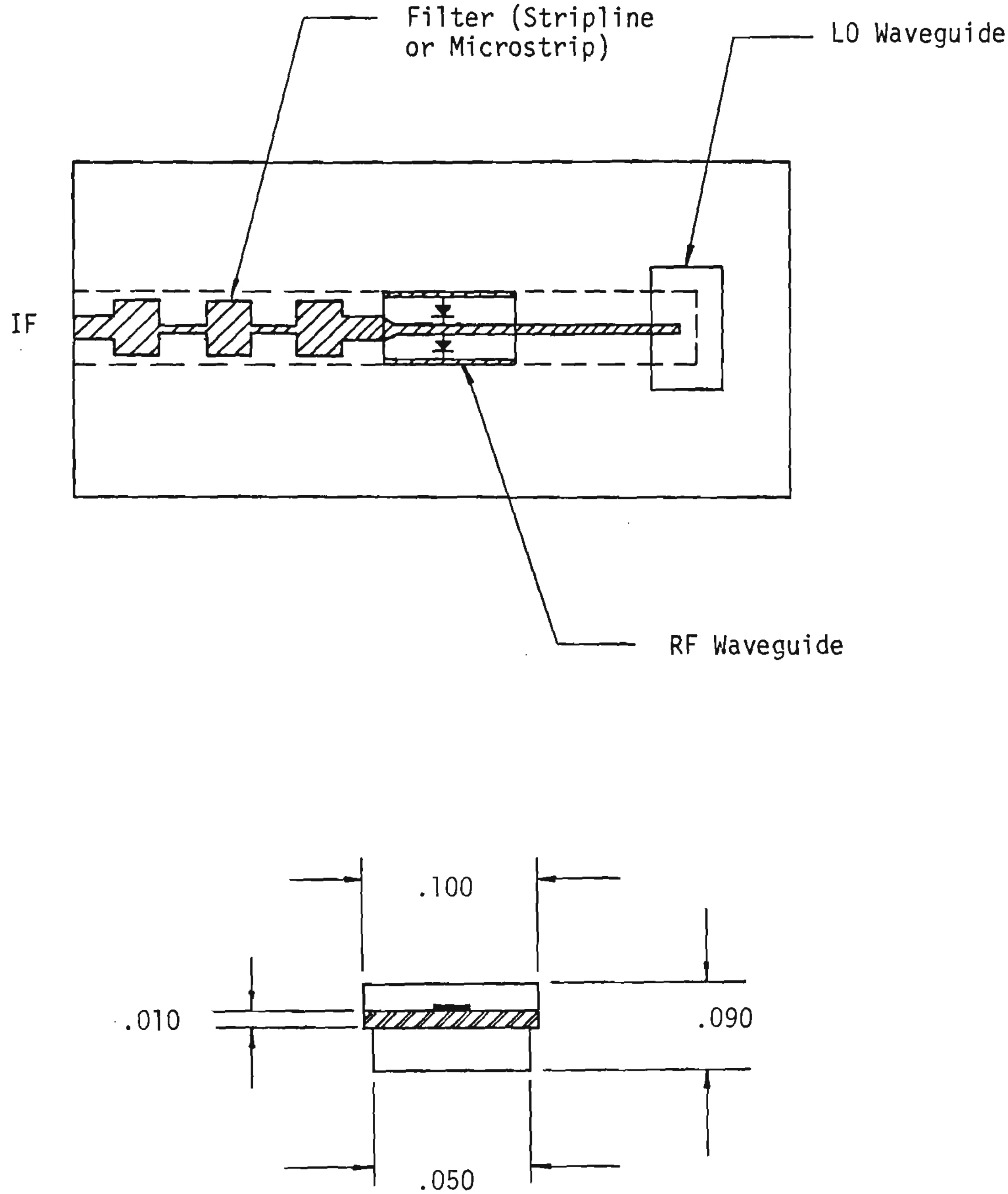


Figure 3-1. Crossbar Mixer Diagram Showing LO/IF H-Plane Feed.

reduce RF losses. The only circuit required is an IF low pass filter and a high pass filter for the LO. The LO can be fed to the diodes by either waveguide, using a waveguide to stripline transition, or directly on a stripline transmission line if a microstrip LO device is available.(3-3, 3-4, 3-5, 3-6, 3-7)

Subharmonic mixers, shown earlier in Figure 2-4, can also be developed in a monolithic fashion with the diodes mounted in antiparallel fashion across the E-plane of an RF waveguide or connected to ground in an enclosed microstrip or suspended substrate stripline channel.

A typical subharmonic mixer diagram is shown in Figure 3-2.

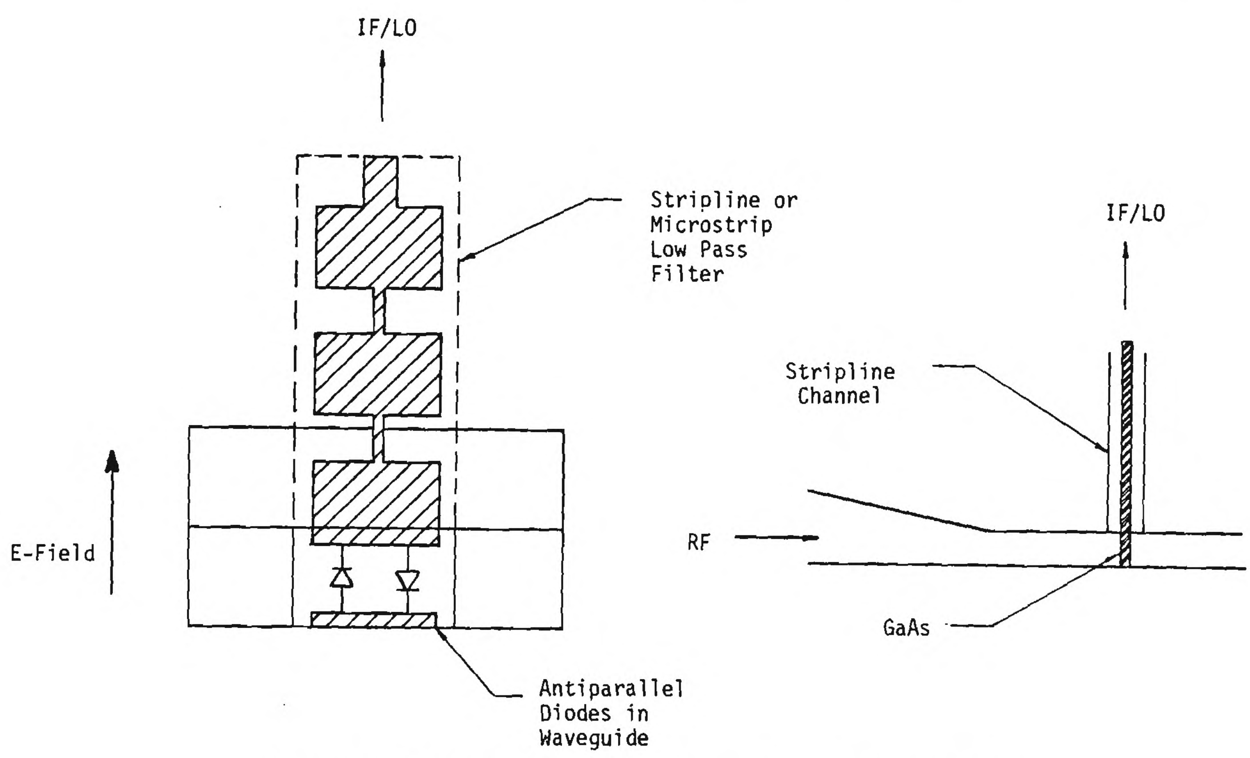


Figure 3-2. Subharmonic Mixer Diagram.

The IF/LO diplexer can be accomplished using waveguide as a high pass filter and a stripline low pass filter which rejects the LO. The low pass filter located just above the diode pair passes the LO but rejects the RF frequency. The IF/LO diplexer can also be made using standard stripline techniques to incorporate monolithic LO and IF amplifier devices. These two diplexing techniques are illustrated in Figure 3-3.

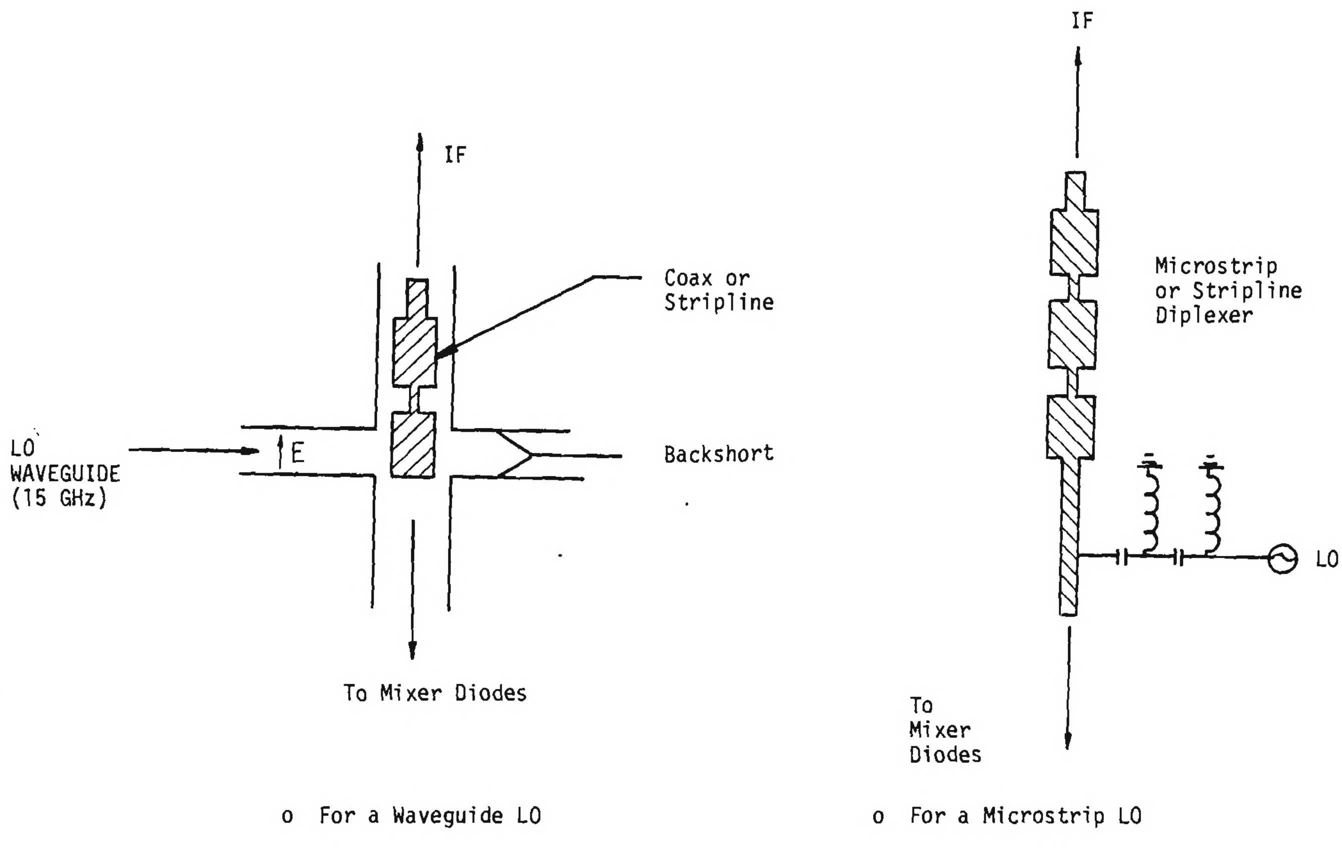


Figure 3-3. Alternate Diplexer Designs.

IV. MIXER CIRCUITS

4.1 Transmission Media and Low Pass Filters

A study of the available transmission line media that are suitable for monolithic receivers has been performed. The findings are that the optimum transmission line media suitable for use in developing Ka -Band GaAs monolithic mixer circuits is either microstrip or suspended substrate stripline. These media, shown in Figure 4-1, can sometimes be conveniently combined with

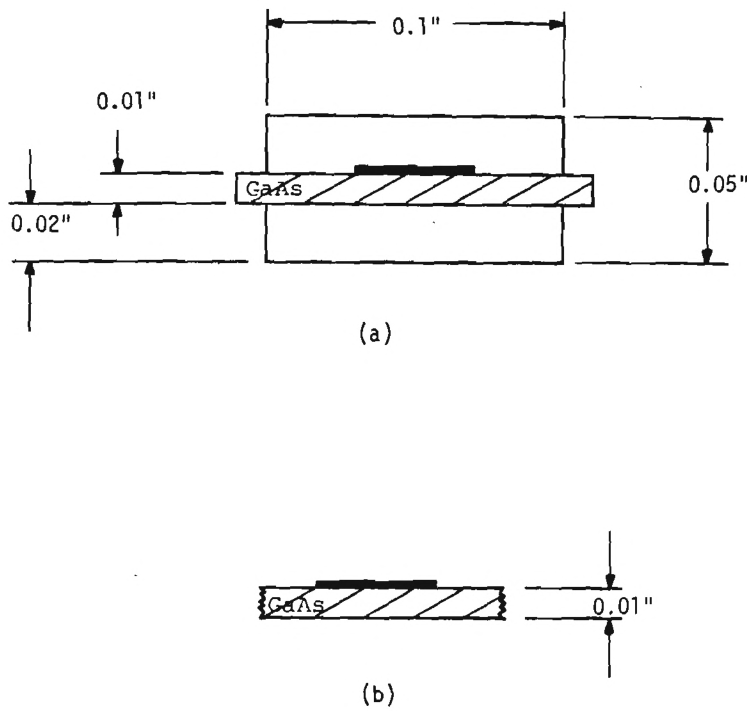


Figure 4-1. Suspended substrate stripline and microstrip cross sections.

waveguide to achieve some of the balanced configurations. Probes can also be used as transitions to waveguide. The final choice of either of these particular media depends on the mixer type, circuits, and coupling to RF, LO, and IF ports. For instance, a mixer that requires a directional coupler, a rat race hybrid, or a directional filter should be made on microstrip in order to reduce circuit size. Some low pass filters, however, would exhibit better performance using suspended substrate stripline because of the higher impedance values that can be obtained with this medium. Suspended substrate stripline would also be preferrable to microstrip for higher frequencies since the air dielectric reduces losses and adds flexibility in choosing transmission line impedances.

A low pass filter is required in almost every mixer type to separate the RF and LO energy (both near 30 GHz) from the IF (at about 7 GHz). Tradeoffs in the design of the low pass filters can be made by: 1) choice of cutoff frequency; 2) number of elements; 3) impedance values of semi-lumped elements; and 4) choice of medium.

Assume, for example, a low pass filter circuit fabricated on a 0.010" thick GaAs substrate. Plots of characteristic impedance and phase velocity for the two preferred transmission line media are given in Figures 4-2 and 4-3. It can be seen from these

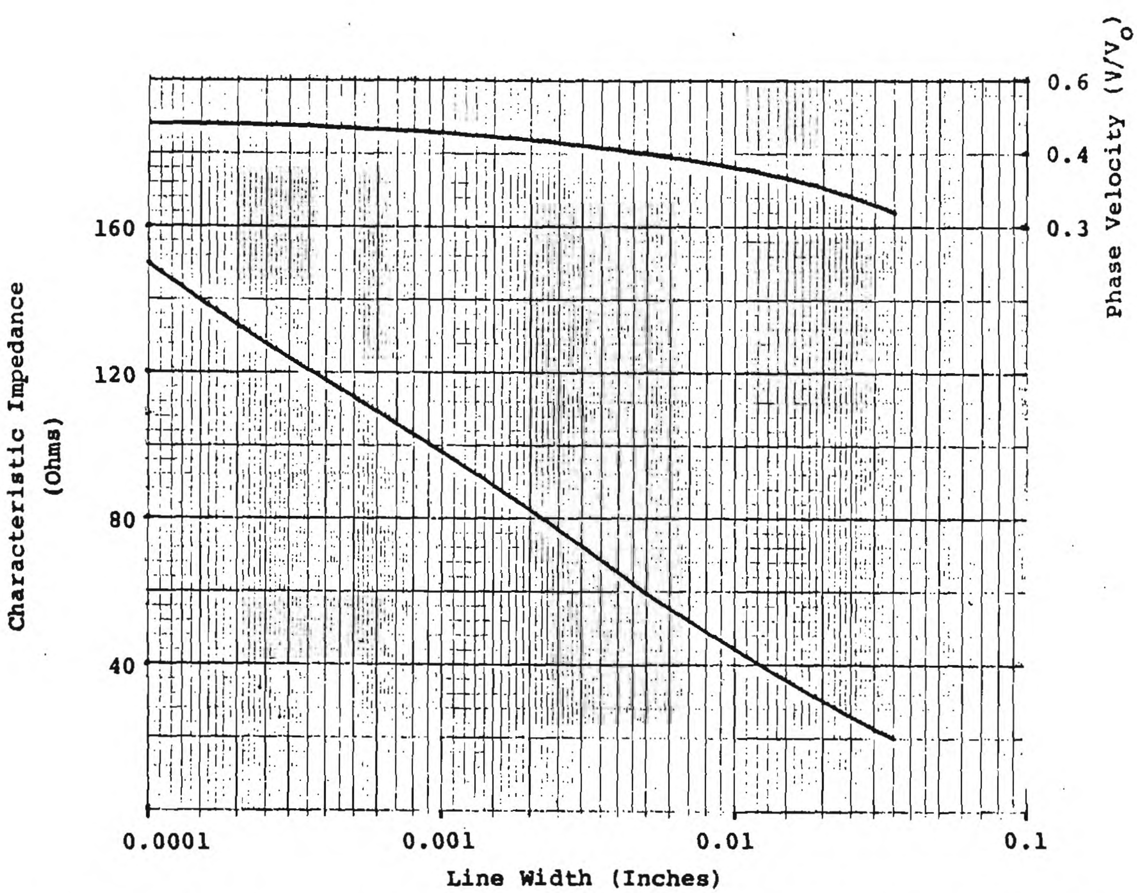


Figure 4-2. Characteristic Impendance and Phase Velocity of 0.01" GaAs Microstrip.

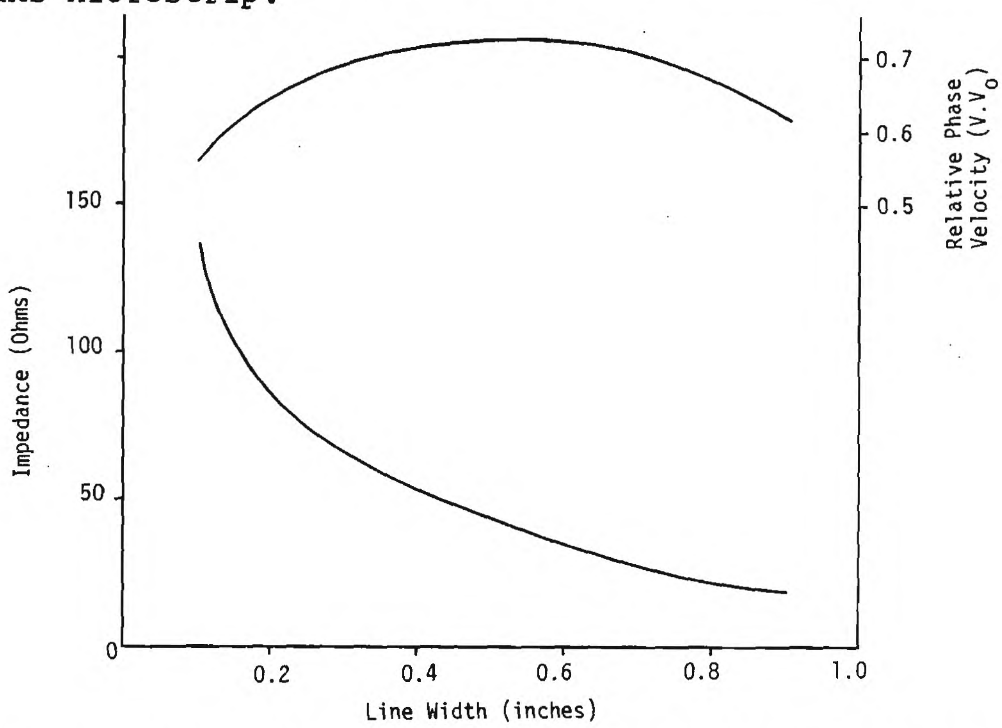


Figure 4-3. Characteristic Impedance and Phase Velocity Versus Line Width for Suspended Substrate Stripline with a 0.01" GaAs substrate.

graphs that higher impedances are more difficult to achieve on microstrip than on suspended substrate stripline. This lowers the values of high impedances reasonably achievable for microstrip. High values are needed to achieve series inductors for the low pass filter. A lower impedance value means that longer lengths of line are needed for a given inductance. An upper limit in length is reached when the length approaches a quarter wavelength at the highest operating frequency. The highest reasonable impedance in microstrip would be about 95 Ohms or a 0.001 inch line width. With suspended substrate stripline, however, 150 Ohm line is achievable. Some filter responses showing the tradeoffs in the filter response for 5 and 7 element filters using either 95 or 150 Ohm inductors are given in Figure 4-4. Overall filter length

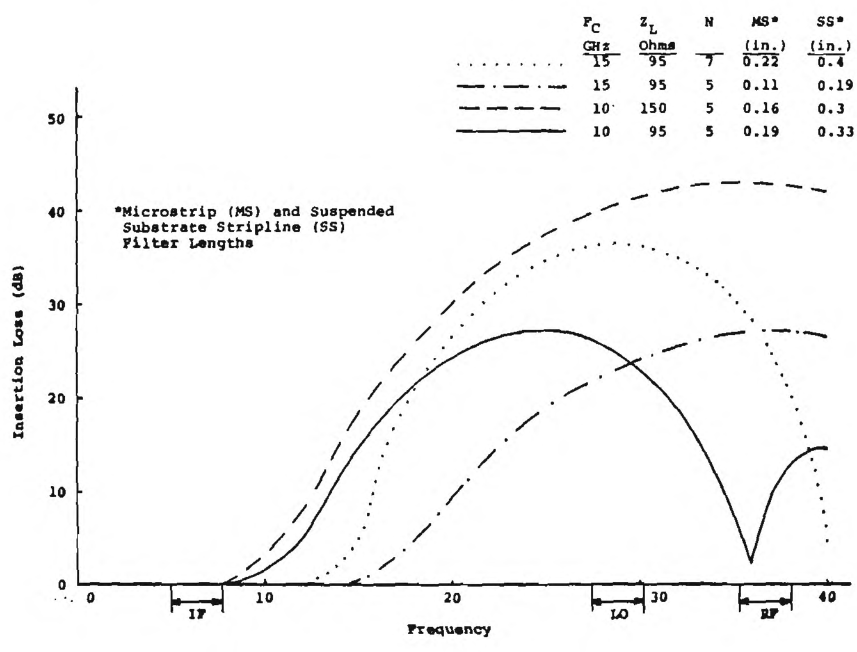


Figure 4-4. Filter Responses for Various Inductor Impedances (Z_L), Cutoff Frequencies (F_C), and Number of Elements (N).

is also provided for both microstrip and stripline versions of these filters. The size of the filter helps determine overall receiver chip size.

4.2 Hybrid Circuits

The rat race hybrid is a 180 degree 3 dB hybrid circuit which splits the LO and RF energy to two diodes as shown in Figure 4-5. The overall size of such a circuit at 36 GHz on 0.01 inch thick GaAs microstrip is shown in the figure.(4-1, 4-2, 4-3)

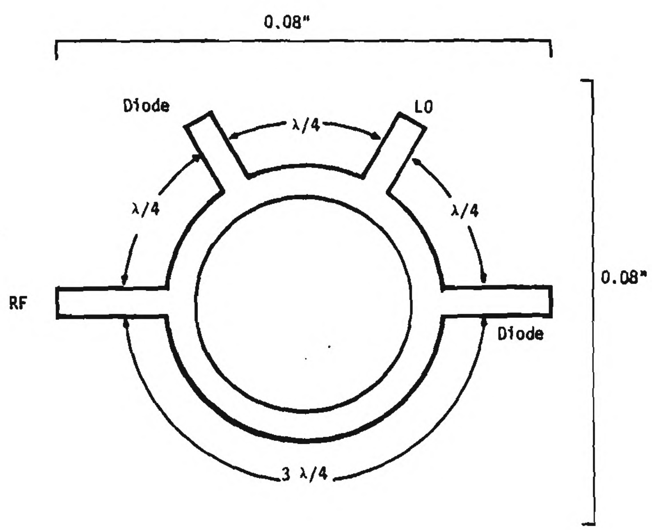


Figure 4-5. Rat Race Hybrid at 36 GHz on 0.01" GaAs Microstrip.

The direct coupled hybrid circuit is a 90 degree coupler and requires an additional length of line for proper phasing of the diodes for balanced mixer operation as does the Lange coupler. Diagrams of these circuits and their overall sizes on GaAs microstrip are shown in Figures 4-6 and 4-7. An estimated 1dB

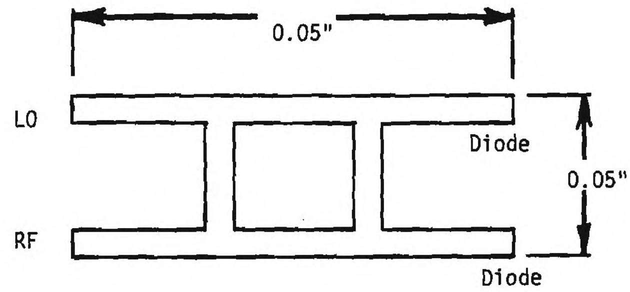


Figure 4-6. Direct Coupled Hybrid on GaAs Microstrip.

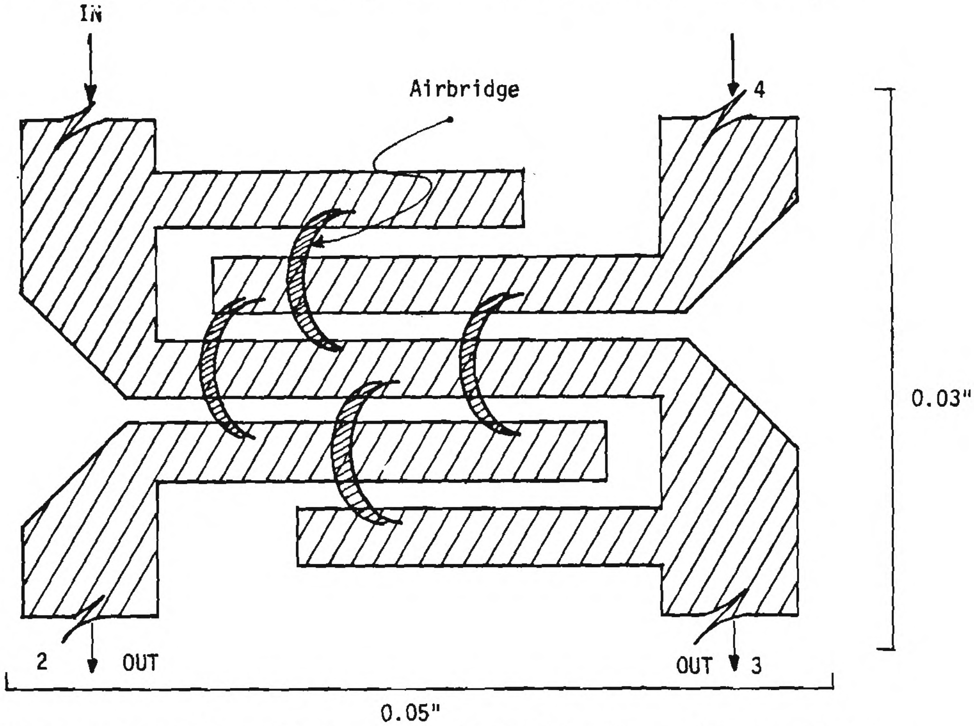


Figure 4-7. Lange Coupler 90 Degree Hybrid Circuit on GaAs.

additional receiver loss is added by the use of these circuits at 36 GHz.[4-8, 4-9]

4.3 Directional Filters and Couplers

These circuits are required for combining the LO and RF energy in a single ended mixer. Diagrams of these circuits for use at about 36 GHz are given in Figure 4-8 and 4-9. The use of

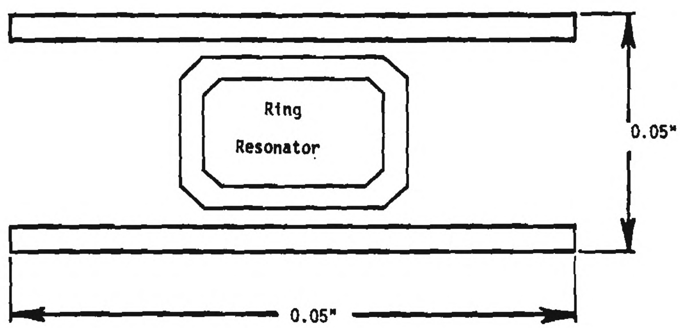


Figure 4-8. GaAs Microstrip Layout of 36 GHz Directional Filter.

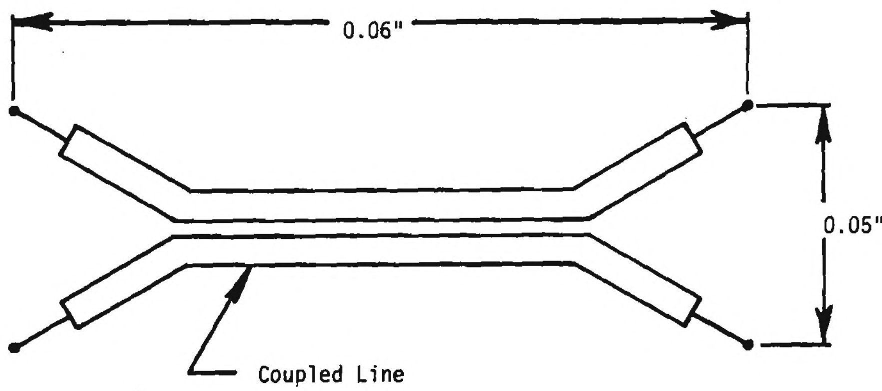


Figure 4-9. 36 GHz Directional Coupler on GaAs.

these circuits to combine the RF and LO also adds about a 1dB additional loss to the receiver.

The directional filter is a four port circuit. It acts as a bandpass filter from the LO to the mixer diode and a bandstop filter from the RF to the mixer diode. About 2-3 dB LO loss is achievable which helps decrease the required LO power when compared to a directional coupler which has about 10 dB LO loss.

The ring resonator portion of the directional filter should be an integral number of wavelengths long at the resonant frequency. Tuning screws are often required to optimize the response of such a microstrip directional filter.

V. MONOLITHIC MIXERS

5.0 Potential Mixer Designs at 36 GHz

Approximate designs of several candidate mixers and circuits for monolithic applications at 36 GHz have been made. This has been done to help evaluate circuit size and interface requirements to other portions of the receiver such as the LO and IF amplifier devices. A 50 Ohm transmission line is available for both the LO and IF ports in each of these designs. All of these circuits are designed for use with 0.01 inch thick GaAs.

Because of the low anticipated LO power (a few milliwatts at 30 GHz) the directional filter approach is suggested for a monolithic single ended mixer design. This will also help reduce LO AM noise because of its bandpass filter characteristics. This design is shown in Figure 5-1 showing the overall circuit dimensions.

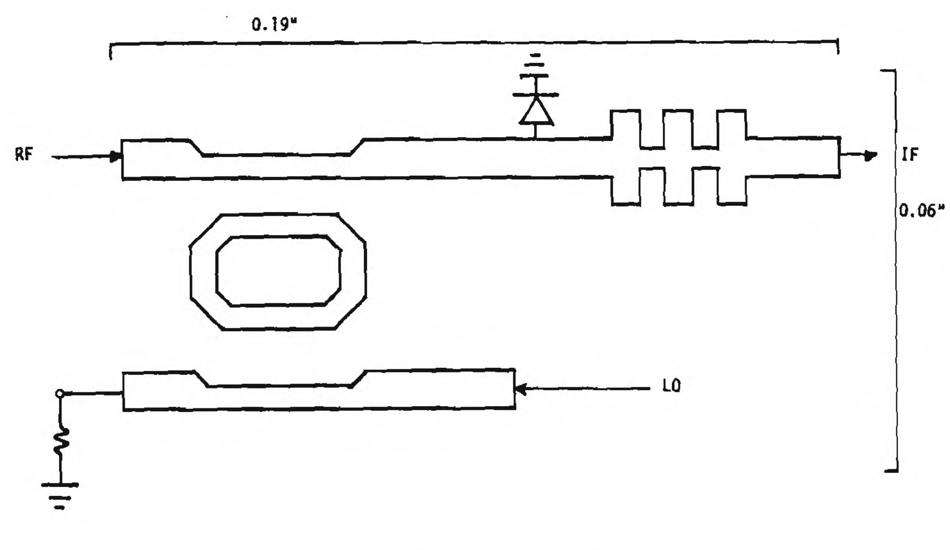


Figure 5-1. Single Ended Monolithic Mixer Layout in GaAs Microstrip.

Layouts of two balanced mixer designs using a Lange coupler and a direct coupled hybrid are given in Figures 5-2 and 5-3. These circuits are quite small and have good potential for high yield because of their size.

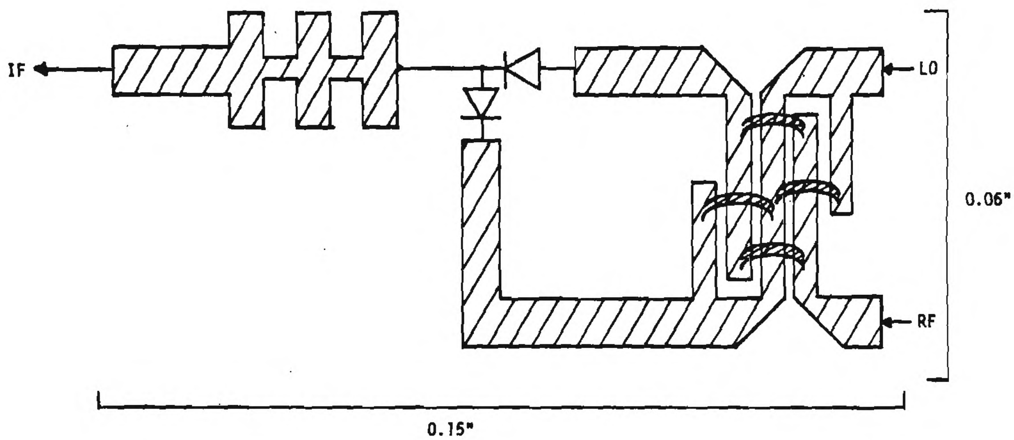


Figure 5-2. Balanced Mixer Layout Using a Lange Hybrid Coupler.

A potential layout for a monolithic crossbar mixer is given is Figure 5-4. This mixer is attractive for monolithic receiver applications because of its simplicity, versatility, and excellent performance as well as its in-line configuration for incorporating microstrip LO and IF devices.

A circuit layout for a subharmonically pumped mixer is given in Figure 5-5. The lowering of the LO frequency makes this circuit extremely attractive for the development of a fully monolithic receiver at 36 GHz.

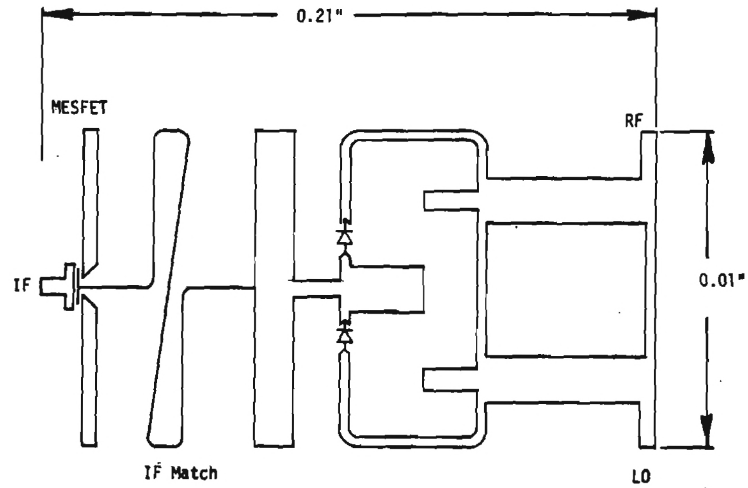


Figure 5-3. Monolithic 31 GHz Receiver Using Direct Coupled Hybrid on GaAs (5-1).

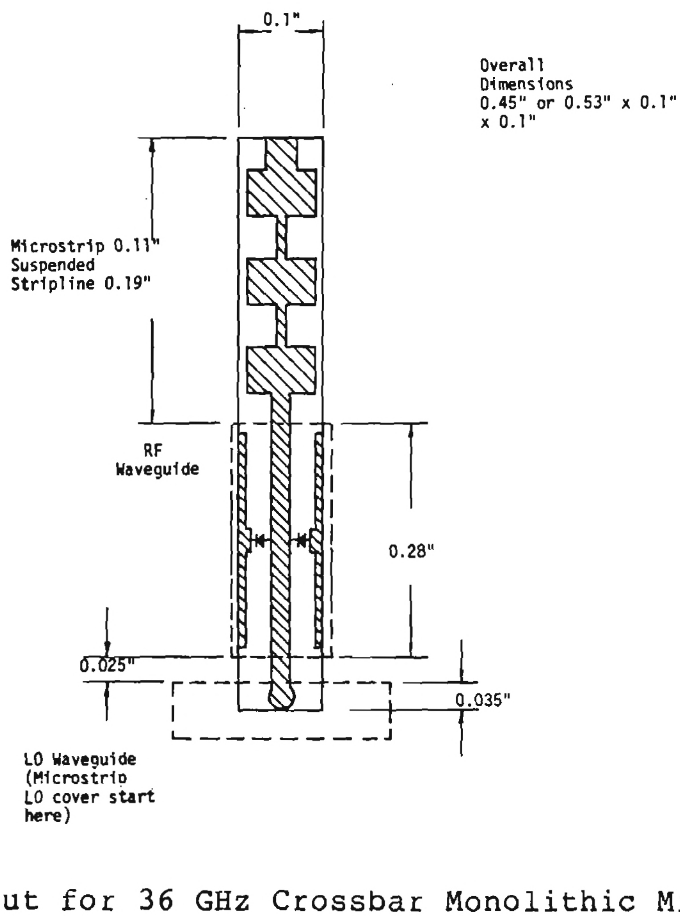


Figure 5-4. Layout for 36 GHz Crossbar Monolithic Mixer.

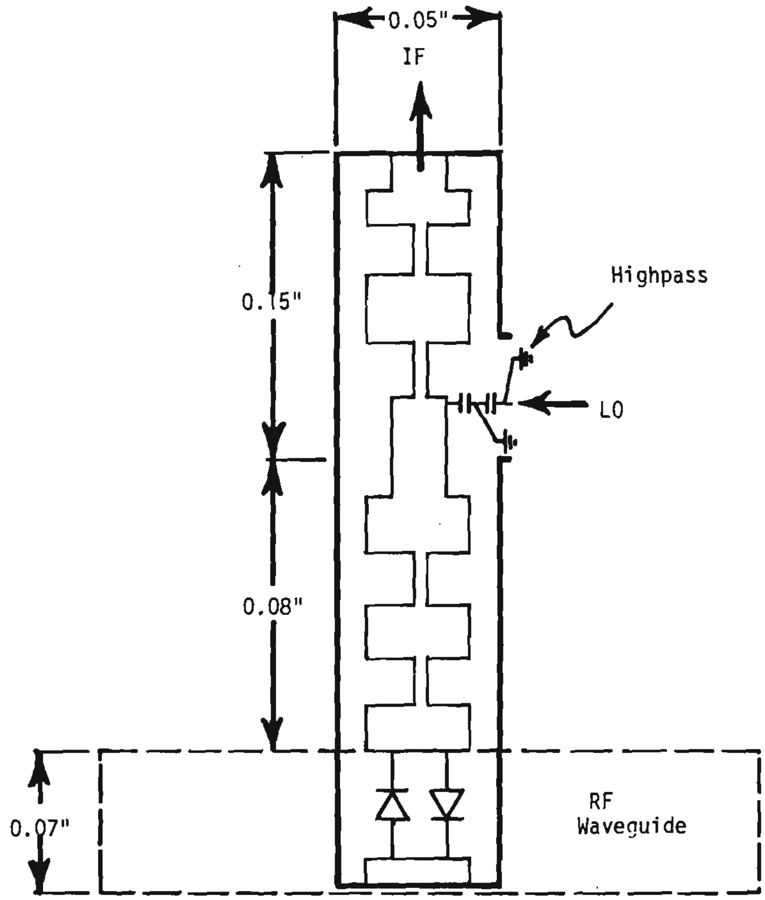


Figure 5-5. Monolithic Subharmonic Mixer Layout on Microstrip.

VI. LOCAL OSCILLATOR APPROACHES

6.1 General Design Approach

The primary candidate devices for integrated local oscillator sources are the FET and Gunn. The FET is routinely fabricated on high resistivity substrates insuring its long term compatibility with mixer diodes and IF technology. Planar Gunn devices have been shown feasible as well, but these devices do not compare well to vertical geometries in terms of either efficiency or phase noise characteristics.

Fundamental frequency (30-35 GHz) FET local oscillators will likely require gate lengths significantly shorter than the half micron standard process devices presently being developed by ITT. Therefore, it is reasonable to consider either a subharmonically pumped approach or a fundamentally pumped approach using a half frequency FET oscillator with doubler. This approach allows for use of half micron or longer devices. If necessary, some initial developmental efforts could employ vertical Gunn device structures integrated in a hybrid fashion. In this instance either the half-frequency or fundamental approach could be tolerated.

With either device, certain critical common problems exist. These include temperature instability, low Q and poor reproducibility. The first two result from the variation of device parameters with temperature and the relatively high noise associated with the low Q oscillator. Questions of yield are more critical in the local oscillator than in any other portion of the receiver. Filter losses, poor mixer diode quality, line width variations, material thickness variation and other factors could all contribute to nominal and gradual performance degradation. But, in the case of the local oscillator, a degree of adjustment must be provided simply to insure that the required frequency can be set.

It is not likely that any integrated monolithic L.O. approach will prove acceptable unless stabilized by some high Q and temperature stable reference means. Possible means include:

1) Phase or frequency locking by loops or injection from a quartz crystal derived stable reference,
2) Cavity stabilization by reaction or transmission techniques using either high Q metal cavities or dielectric resonators,
3) automatic frequency control (stabilization) by virtue of varactor tuning and a high Q discriminator network or
4) open loop frequency control of a varactor tuned oscillator using a temperature sensor and PROM bias control to retain the nominal frequency constant as temperature varies. No noise reduction results from this approach. It is a means, however, of reducing the frequency excursion over which other stabilizing means must be well behaved.

Each of these approaches offers its own set of advantages and disadvantages. Consider first oscillators that employ electronic tuning and a stable reference to obtain 1) midband frequency setting, 2) FM noise reduction and 3) temperature stabilization.

6.2 <u>Electronically Tunable Oscillators</u>

A) FET Doubling Oscillator

Using 1 micron NEC 69400 FETs, Winch(6-1) has reported a FET doubling oscillator tunable over the 19.5 to 30.6 GHz range. His schematic and performance are given in Figure 6-1.

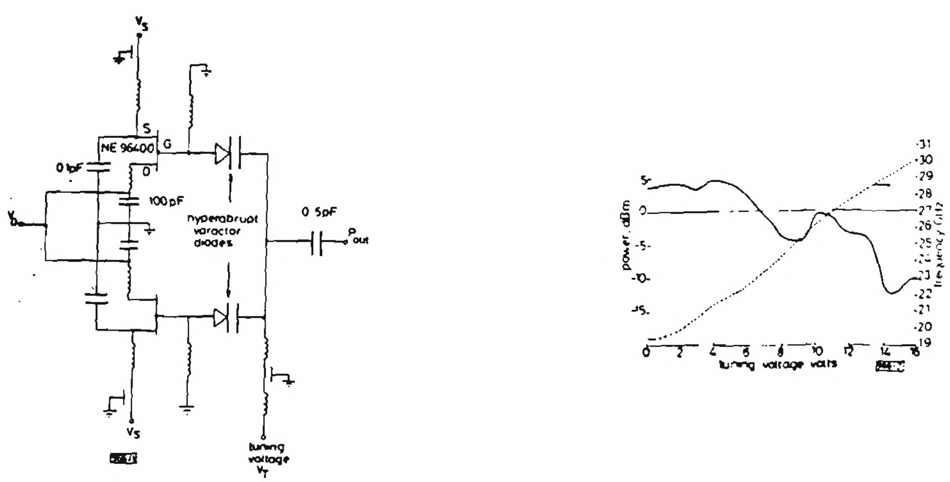


Figure 6-1. FET Doubling Oscillator (Winch(6-1))

The broad tuning range of this type of oscillator is a distinct advantage in that the same nominal design could prove useful for more than one application. Another advantage of such a broad tuning range is the ability to compensate readily for temperature frequency variations. Finally, by use of the doubling approach, FET requirements are far less stringent. A disadvantage is the requirement for the hyperabrupt varactors. This disadvantage is, however, no worse than that incurred with any half frequency source followed by a doubler. Another disadvantage is the power absorption in the varactor(s) when large tuning ranges are sought. The power roll-off problem must be solved regardless of the approach.

B) Monolithic Doubler

Chu et al[6-2] have recently reported a monolithic doubler employing planar Schottky barrier varactors. Their circuit provided 20-25% doubling efficiency with power output of about 20 mW at frequencies up to 27.4 GHz. Their circuit (Figure 6-2) has dimensions of 4 x 8 mm (.160 x .320 inch). No particular effort was made to fold or reduce the size of any of the circuit elements. With higher frequency of 35 GHz and other efforts to reduce size, it should be possible to fabricate the circuit in a region approximately 2.5 by 5 mm (.100 x .200 inch). The bandwidth achieved with this doubler indicates its use with an appropriate tunable oscillator.

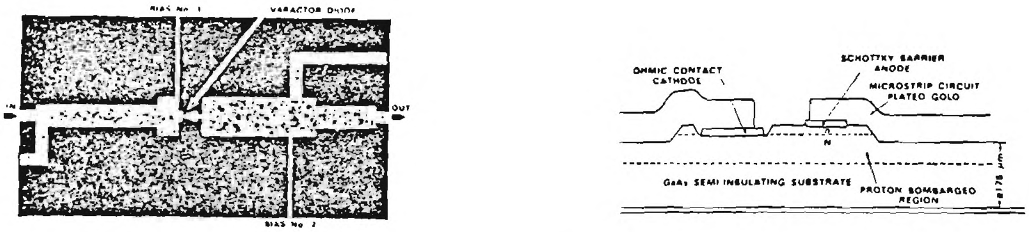


Fig. 6-2 FET Doubler (Chu et al[6-2]), Circuit and Planar Schottky Varactor

C) FET Doubling

Rauscher[6-3] has reported both FET doublers and FET doubling oscillators using Avantek M106 chips. He predicted and achieved conversion gain of -1dB for an input level of +1dBm at 30 GHz in a FET doubler. His doubling oscillator provided 8.8 dBm of output at 29 GHz with 8.9 percent efficiency.

These three examples (A, B, C) appear to indicate that FET or Varactor doublers or FET doubling oscillators can provide the necessary tunability and power output levels to drive fundamental frequency mixers in Ka-band. Discussions with R. Trew indicate that a half frequency oscillator with doubler could be a desirable approach with respect to noise performance.

D) Tunable Gunn Oscillator

Rubin(6-4) has reported an example of a typical varactor tuned Gunn oscillator for 35 GHz (Figure 6-3). His oscillator provided approximately 5 milliwatts of output power over a 1 GHz range as noted in Figure 6-3.

The oscillator is fabricated on a low dielectric constant substrate ($\varepsilon = 2.3$) and has dimensions of about .90 x .75 inch including an extensive ground plane. Similar results could be achieved, using different grounding techniques, on a substrate

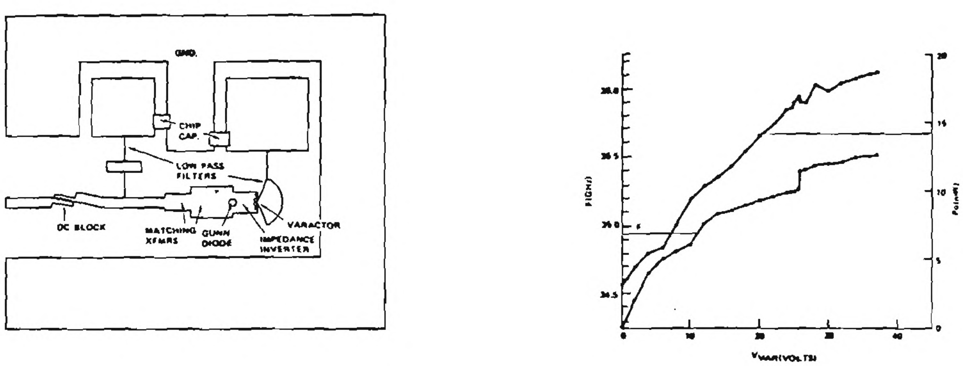


Figure 6-3. Varactor Tuned Gunn Oscillator (Rubin(6-4))

about .75 x .50 inch. With gallium arsenide substrates such a design would reduce to approximately .3 x .2 inch, still a rather large size for integration with the remaining L.O. and receiver elements.

Given the foregoing examples (many other are available) it is reasonable to take the view that a tunable local oscillator can be developed. If so, several possible means of stabilization and/or FM noise reduction exist.

6.3 Frequency Control/Locking

A) General Comment

It is usually necessary, where phase locked arrangements are used, to specify a range of sideband frequencies over which stabilization is required. The complexity of the stable reference and feedback loop are greatly affected by these specific requirements. In the next few paragraphs no attempt is made to consider any specific example. Instead, several nominal approaches that may prove useful are discussed.

B) Subharmonic Injection Locking

It has been shown that signals injected at a subharmonic of the output frequency can give results similar to that obtained with fundamental frequency locking. Unfortunately, locking gain and bandwidth are both quite low for this scheme. Its use is not recommended.

C) Parametric Injection Locking (PIL)

Okamoto[6-5] has demonstrated a reasonable circuit for parametric injection locking using an IMPATT operated at 75 GHz. He achieved noise reduction of about 25 db (as noted on a spectrum analyzer) and frequency stability as good as $10^{-8}/^{o}C$. Locking range of 700 MHz (1%) was obtained with injected power of 23 dBm. His circuit is shown schematically in figure 6-4.

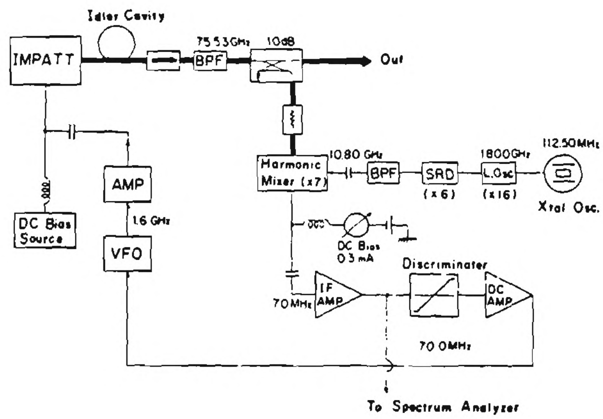


Figure 6-4. Parametric Injection Locked Oscillator Schematic (OKAMOTO(6-5))

Parametric injection locking offers the advantage that a low-frequency stable-oscillator can be employed for locking. A significant disadvantage is that a high Q idler resonator, operating near the output frequency, is also required to achieve Okamoto's results. Furthermore, the stability and/or noise reduction depend on both the cavity and the stable low frequency signal. This approach is an unlikely choice at present.

D) Modified Pound[6-6] Circuit

Chan and Cole[6-7] have reported a stabilized Gunn oscillator in X-band using a circuit similar to that presented by Pound[6-6] in 1946. At X-band, they achieved an average short term stability of 2 x 10^{-7} and temperature stability of 9 x $10^{-7}/^{o}C$ between -15 and +30^{o}C. The nominal oscillator and electronic circuits are shown in figure 6-5 for reference purposes only. We do not suggest that such a complex approach should be attempted at present. However, for sophisticated requirements, this level of complexity may be required. For less sophisticated requirements a relatively simple circuit such as that discussed next may suffice.

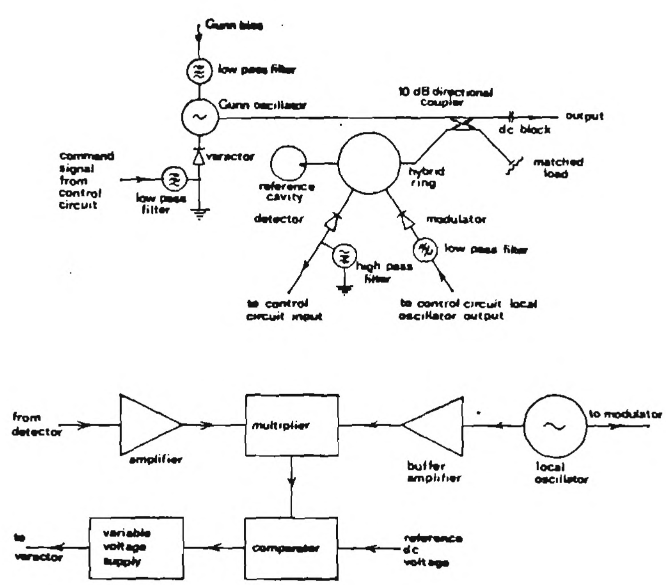


Figure 6-5. Modified Pound Circuit (Chan[6-7])

E) Discriminator Stabilized Oscillator

Glance and Snell(6-8) have described a very simple approach to a microstrip discriminator for use in C-band (figure 6-6). The circuit is just the usual S - curve discriminator achieved by use of a microstrip resonator having unequal rectangular dimensions. When coupled with a varactor tunable Gunn oscillator, this discriminator provided a stabilization factor of 1000 up to 1 KHz where the I.C. amplifier gain response began to decrease. The stabilized oscillator, including the I.C. amplifier, is shown in figure 6-7 along with the response characteristics of the discriminator.

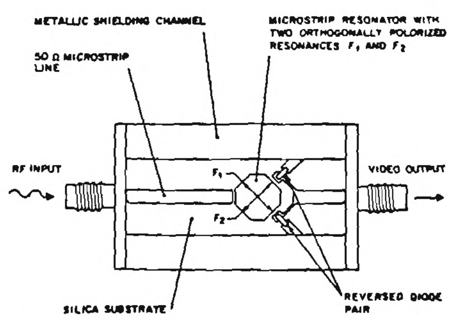


Figure 6-6. S-Curve Microstrip Discriminator

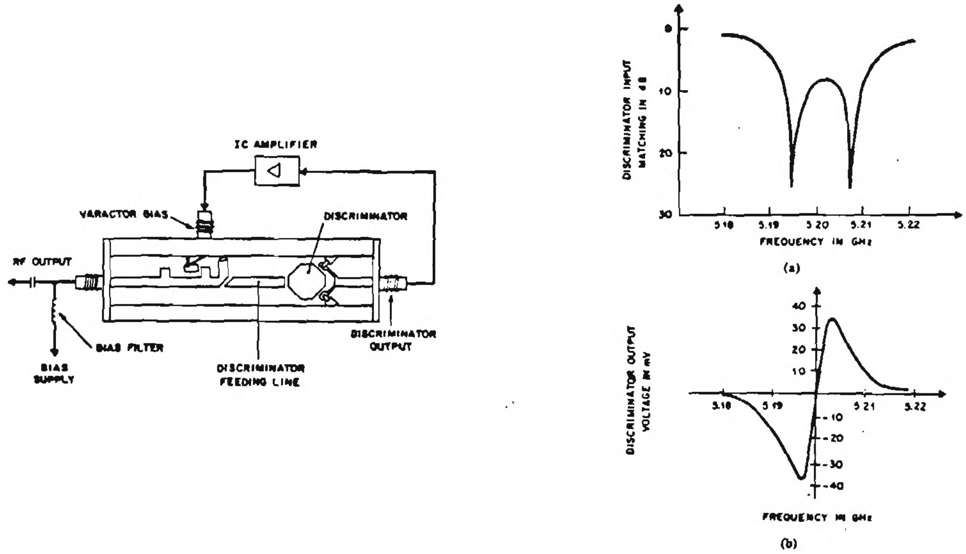


Figure 6-7. Stabilized Oscillator and Discriminator Characteristics.

In this particular example, the maximum voltage of the I.C. was limited to a saturated value which would never cause the oscillator to tune outside the capture limits of the S-curve and saturated amplifier combination. Therefore no search mode was required within the closed loop. For this simple approach to be useful, the total frequency excursion of the basic low Q oscillator, due to manufacturing tolerances, temperature or any other effect, must not exceed the range of frequencies over which the amplifier's output is saturated. With reduced S-curve slope and more complex feedback electronics, even this range could be extended.

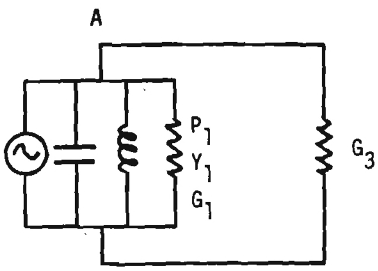
Given the simplicity of this basic approach, it is interesting to consider use of a discriminator employing an asymmetric rectangular dielectric resonator to achieve either a K_u or K_a-band discrminator. This discriminator could be coupled readily with any of the electronically tuned oscillators already discussed. An ideal combination might ultimately prove to be the discriminator coupled with the FET doubling oscillator suggested in paragraph 6.2.A or with a varactor tuned half frequency L.O. near 15 GHz.

Fixed frequency dielectric resonator stabilized oscillators are discussed next. It is interesting to note here, however, that the discriminator scheme might allow one to first prestabilize the oscillator using a fixed tuned dielectric resonator and then to further stabilize in selected IF bands by use of the discriminator and a band limited feedback amplifier.

6.4 OSCILLATOR STABILIZATION WITH HIGH Q RESONATOR

6.4.1 Basic Concept

The basic concept of passive stabilization is an unusually simple one. By way of illustration, consider the elementary signal source, A, of figure 6-8.



P_1 = max. available power

Y_1 = Admittance at P_1

Figure 6-8 Elemental Source

The Q of the source when loaded for maximum power is:

$$Q_{L1} = \left| \frac{\omega_o \frac{\partial B}{\partial \omega}}{2G_1} \right|_{\omega=\omega_o} = \frac{B_1}{2G_1} \triangleq \frac{Q_1}{2}$$

where $G_3 = G_1$.

If another resonant circuit is added as shown in figure 6-9,

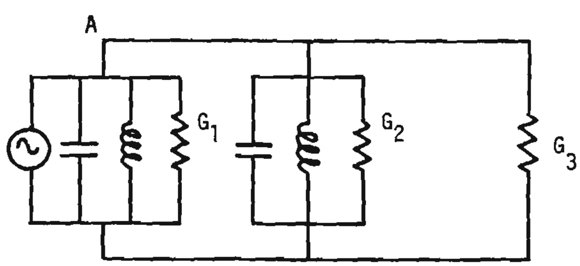


Figure 6-9. Elemental Source (A) with Additional Resonator and Load.

the admittance is:

$Y_T = G_1 + G_2 + G_3 + j(B_1 + B_2)$

where $G_1 = G_2 + G_3$ for matched loading.

Then the overall loaded Q, Q_{LT}, is:

$$Q_{LT} = \frac{B_1+B_2}{G_1+G_2+G_3} = \frac{B_1+B_2}{2G_1} = \frac{B_1}{2G_1} + \frac{B_2}{2G_1} + \frac{B_2}{2G_2}\left(\frac{G_2}{G_1}\right)$$

$$= \frac{Q_1}{2} + \frac{Q_2}{2}\left(\frac{G_2}{G_1}\right) = \frac{Q_1}{2} + \frac{Q_2}{2}\left(1 - \frac{G_3}{G_1}\right)$$

$$Q_{LT} = \frac{Q_1}{2} + \frac{Q_2}{2}\left(1 - \frac{P_3}{P_1}\right) \qquad (2)$$

wnere P_1 is the maximum available power and P_3 is the power transmitted to the load. P_2, then, is just the portion of the available power dissipated in the resonator.

If we wish to define a factor S showing the improvement in Q caused by adding the second (or stabilizing) resonance we may set

$$S \triangleq \frac{Q_{LT}}{Q_{L1}}$$

such that

$$S = 1 + \frac{Q_2}{Q_1}\left(1 - \frac{P_3}{P_1}\right) \qquad (3)$$

wnere $Q_1 = \frac{fo}{2.4L}$ and L is the conventional pulling factor defined as the total frequency excursion of the source (A) alone, connected as in figure 6-8, when the load is varied through all phases of a 1.5:1 VSWR.

Thus, the factor S is the conventional "S-factor" used throughout the industry to specify the reduction of frequency fluctuations by passive stabilization.

6.4.2 Basic Concept with Adjustable Coupling

In the foregoing case, no means was provided to adjust the relative power split between the resonator and the load; therefore, for high Q_2 where $G_2 << G_3$ and where $G_2 \rightarrow G_1$ to retain maximum power,

$$S \rightarrow 1 + \frac{dB_2}{dB_1} \quad \text{and} \quad \frac{P_3}{P_0} \rightarrow 0$$

This is just the limiting case where all or nearly all of the available power is dissipated in the stabilizing cavity, thus leading to maximum stabilization and minimum power delivered to the load.

If means are included to adjust the relative dissipation in the resonator and load, the circuit of figure 6-10 results:

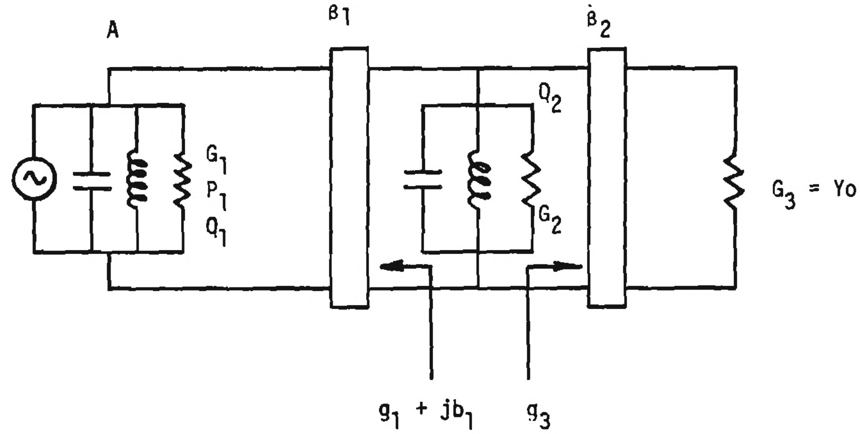


Figure 6-10. Stabilizer with Adjustable Coupling

where $\beta_1 \triangleq \frac{g_1}{G_2}$ and $\beta_2 \triangleq \frac{g_3}{G_2}$.

But to obtain maximum power from the source,

$$g_1 = G_2 + g_3 \text{ such that } \beta_1 = 1 + \beta_2$$

Once more determining $S = \frac{Q_{LT}}{Q_{L1}}$, obtain:

$$Q_{LT} = \frac{b_1+B_2}{g_1+G_2+g_3} = \frac{Q_1}{2} + \frac{B_2}{2g_1} = \frac{Q_1}{2} + \frac{B_2G_2}{2G_2g_1}$$

$$= \frac{Q_1}{2} + \frac{Q_2}{2}\frac{P_2}{P_1} = \frac{Q_1}{2} + \frac{Q_2}{2}\left(1 - \frac{P_3}{P_1}\right)$$

$$\text{and } S = 1 + \frac{Q_2}{Q_1}\left(1 - \frac{P_3}{P_1}\right), \tag{4}$$

again, the conventional "S-factor".

In the matched case being considered,

$$\frac{P_3}{P_1} = \frac{\beta_2}{\beta_1} = \frac{\beta_1-1}{\beta_1} = 1 - \frac{1}{\beta_1} ,$$

which is just the efficiency with which the maximum available power is transferred to the load.

6.4.3 Transmission Line Case

Consider next the more complicated case of a microwave oscillator connected through a transmission line to a transmission cavity or resonator and through the resonator to a load. See figure 6-11.

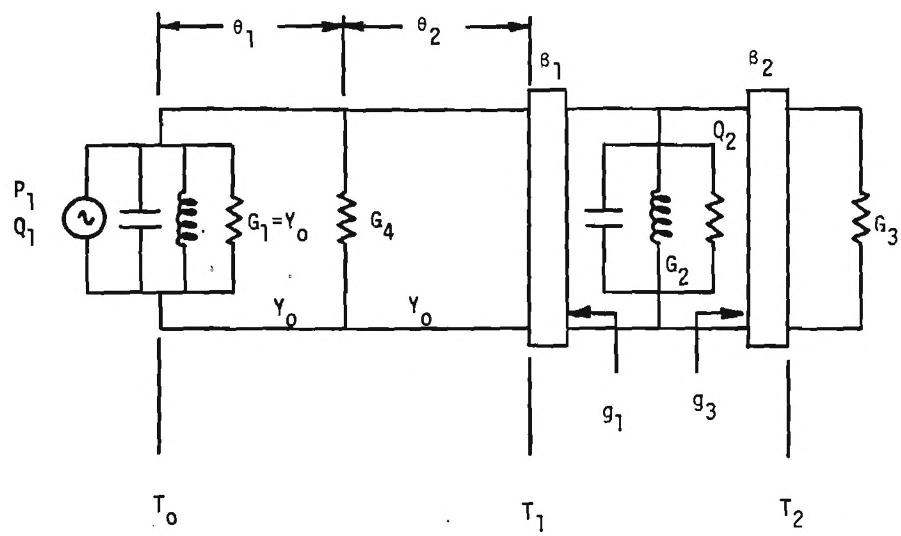


Figure 6-11. Transmission Line Coupled Stabilizer Circuit

The conductance G_4 is a mode stabilizer which is usually required to prevent frequency hops and lock-up at one or more low Q resonances associated with the various reactive coupling elements and transmission line segments θ_1 and θ_2. Achievement of high values of S almost always requires this stabilizer.

It is almost always desired that the basic oscillator see matched load conditions (maximum power) requiring that

$$g_1 = G_2 + g_3 \ .$$

But, $g_1 = \beta_1 G_2$ and $g_3 = \beta_2 G_2$, giving

$$\beta_1 = 1 + \beta_2 \tag{5}$$

and
$$\beta_2 = \frac{g_3}{G_2} = \frac{P_3}{P_2} = \frac{P_3}{P_1 - P_4 - P_3} \ .$$

Therefore,
$$\beta_1 = \frac{P_1 - P_4}{P_1 - P_4 - P_3}$$

The circuit of Figure 6-11 simplifies to that of Figure 6-12 by transforming the oscillator and load to reference planes T_1 and T_2 respectively.

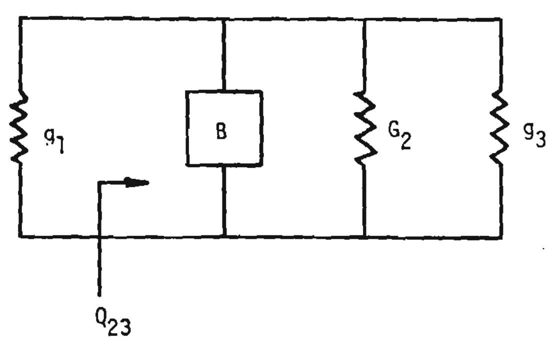


Figure 6-12. Circuit Representation of the Stabilizer Resonator with its Loads.

For any general circuit:

$$B_{Total} = B_1 + B_2 + B_3 + \cdots + B_j$$

and $$B_j = Q_j G_j$$

Therefore,

$$Q_{23} = \frac{G_2}{G_2+g_3} \cdot Q_2 = \frac{G_2}{G_2+\beta_2 G_2} \cdot Q_2$$

$$Q_{23} = \frac{1}{1+\beta_2} \cdot Q_2 = Q_2 \left(1 - \frac{P_3}{P_1-P_4}\right) . \qquad (6)$$

Eq. (6) is obtained by substitution of the right side of Eq. (5) into β_2, and noting that

$$G_T Q_T = Q_1 G_1 + Q_2 G_2 + \cdots + Q_j G_j$$

Therefore, the overall loaded Q of the stabilized oscillator is:

$$Q_L = \frac{Q_1 g_1 + Q_{23}\,(G_2+g_3)}{g_1+G_2+g_3} + \frac{Q_1 g_1 + g_1 Q_{23}}{2g_1}$$

$$Q_L = \frac{Q_1 + Q_{23}}{2}$$

$$\text{Therefore,} \qquad S = \frac{Q_L}{Q_{1/2}} = \frac{Q_1 + Q_{23}}{Q_1} = 1 + \frac{Q_{23}}{Q_1}$$

$$\text{or} \qquad S = 1 + \frac{Q_2}{Q_1}\left(1 - \frac{P_3}{P_1 - P_4}\right)$$

$$\text{and} \qquad S = 1 + \frac{Q_2}{Q_1}\left(1 - \frac{P_3}{1 - \frac{P_4}{P_1} \quad P_1}\right)$$

For simplicity, assume that the admittance of the source is adjusted slightly to compensate for the effect of G_4 on the match, such that:

$$Y_o'\Big|_{source} = \frac{Y_o^2}{G_4+y_o} \quad , \quad \frac{Y_o'}{Y_o} = \frac{1}{1 + \frac{G_4}{y_o}}$$

The Q of the source, Q_1, is not affected by this change.

Then,

$$\frac{P_4}{P_1} = \frac{\frac{G_4}{Y_o}}{1 + \frac{G_4}{y_o}}$$

and $$S = 1 + \frac{Q_2}{Q_1}\left(1 - \frac{P_3}{P_1}\left(1 + \frac{G_4}{y_0}\right)\right) \tag{7}$$

This same simplification can also be realized by altering the characteristic admittance of the first quarter wave line section such that:

$$Y_0'\Big|_{line} = \sqrt{y_0(G_4 + y_0)} \quad ;$$

again, a small change having no effect on Q_1.

Note also that in some cases it is desirable to estimate the absolute value of the loaded stabilized Q rather than compare the stabilized Q with that of the unstabilized Q. The overall loaded stabilized Q is given by:

$$Q_{LT} = S \cdot \frac{Q_1}{Q_2}$$

$$= \frac{Q_1}{2} + \frac{Q_2}{2}\left(1 - \frac{P_3}{P_1}\left(1 + \frac{G_4}{y_0}\right)\right) ,$$

subject to the same simplifying modifications noted above.
The stabilization factor of (7) is the same as that already derived except for the effect of the stabilizer G_4. This effect is almost always ignored in the calculation of S. Unfortunately, with solid state oscillators $^{G_4}/Y_0$ can often be 0.2 or greater and

its effect cannot be ignored.

Consider a typical case where:

$$Q_2 = 5000$$

$$Q_1 = 50$$

$$P_3 = .5\ P_1$$

$$\frac{G_4}{Y_o} = .15, \text{ a typical value.}$$

Then, $S = 42.5$

But for $\frac{G_4}{Y_o} = 0$

$S = 51.$

Clearly, the effect of the stabilizer on S is not trivial; the predicted factors differ by about 20 percent. Unfortunately, $\frac{G_4}{Y_o}$ cannot usually be kept small as is evident from the discussion which follows.

6.4.4 Stability Considerations

The load admittance presented to the basic oscillator (at terminals T_o of figure 6-11) is given by:

$$\frac{Y}{Y_o} \quad \frac{A + j \tan\left[(2n-1)\frac{\pi}{2}\cdot\frac{f}{f_o}\right]}{1 + jA \tan\left[(2n-1)\frac{\pi}{2}\cdot\frac{f}{f_o}\right]} \qquad (8)$$

where
$$A = \frac{G_4}{Y_o} + \frac{1 + j\,Q_{23}\left(\frac{f}{f_o} - \frac{f_o}{f}\right) + j\tan\left(\frac{\pi}{2}\cdot\frac{f}{f_o}\right)}{1 + j\left[1 + j\,Q_{23}\left(\frac{f}{f_o} - \frac{f_o}{f}\right)\tan\left(\frac{\pi}{2}\cdot\frac{f}{f_o}\right)\right]}$$

and
$$Q_{23} = \left(1 - \frac{P_3}{P_1}\left(1 + \frac{G_4}{y_o}\right)\right) Q_2$$

Consider the earlier case where $Q_2 = 5000$ and require that half the available power P_1 be delivered to the load, giving $\frac{P_3}{P_1} = .5$. If we now plot the real and imaginary parts of (8) for various values of stabilizing resistor G_4, figures 6-13 and 6-14, it becomes obvious why significant values of G_4 must be employed. If we assume that the susceptance slope of the elemental source is constant over the relatively narrow band of the plots we can plot $-\frac{B_1}{Y_0}$ on the B/Y_0 plot of figure 6-13. Five intersections with the B/Y_0 graph can occur. Of these, the second and fourth cannot result in stable oscillations because $\partial B/Y_0/\partial\omega < 0$. Points 1, 3 and 5, on the other hand, have $\partial B/Y_0/\partial\omega > 0$ and may, under some bias condition of start-up, constitute the operating point if $G_1/Y_0 = G/Y_0$. The third point ($B/Y_0 = 0$, $f/f_0 = 1$) is the only point where the predicted and high S-factor will be achieved. Therefore, points 1 and 5 must somehow be eliminated. This is accomplished by increasing the stabilizing conductance G_4/Y_0 until no undesired intersection occurs. Note that for $Q_1 = 31.25$ and $G_4/Y_0 = .200$ the curves are tangential implying a very low Q and noisy oscillation potential near the frequency of the tangency. For slightly higher Q_1 or larger G_4/Y_0, intersections 1 and 5 are eliminated leading to stable conditions only at point 3, the desired operating point.

NOTE 1): It is more common to describe the foregoing stability considerations by use of either G-B plots or Smith chart plots. Then, the load admittance and its intersection

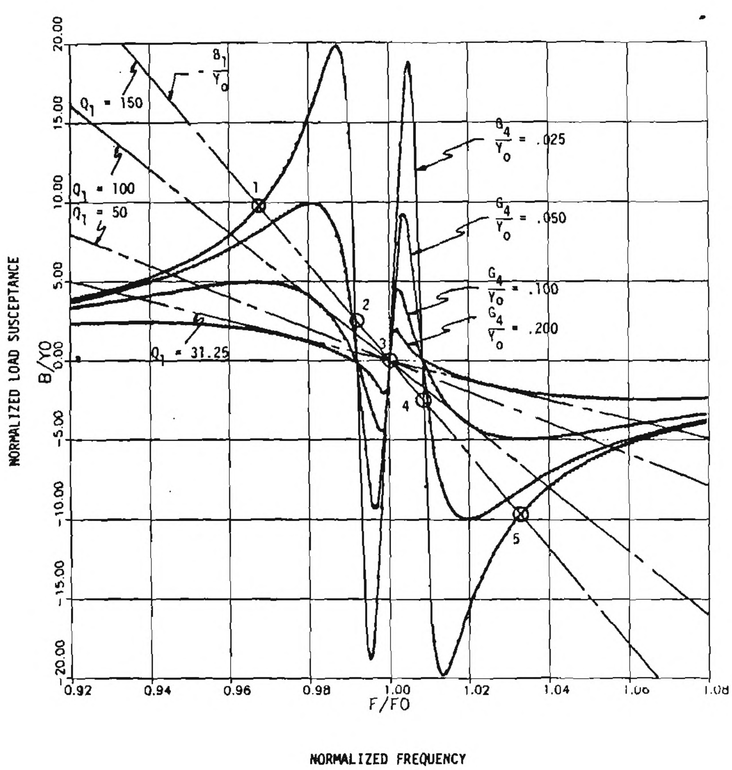


Figure 6-13. Load Susceptance.

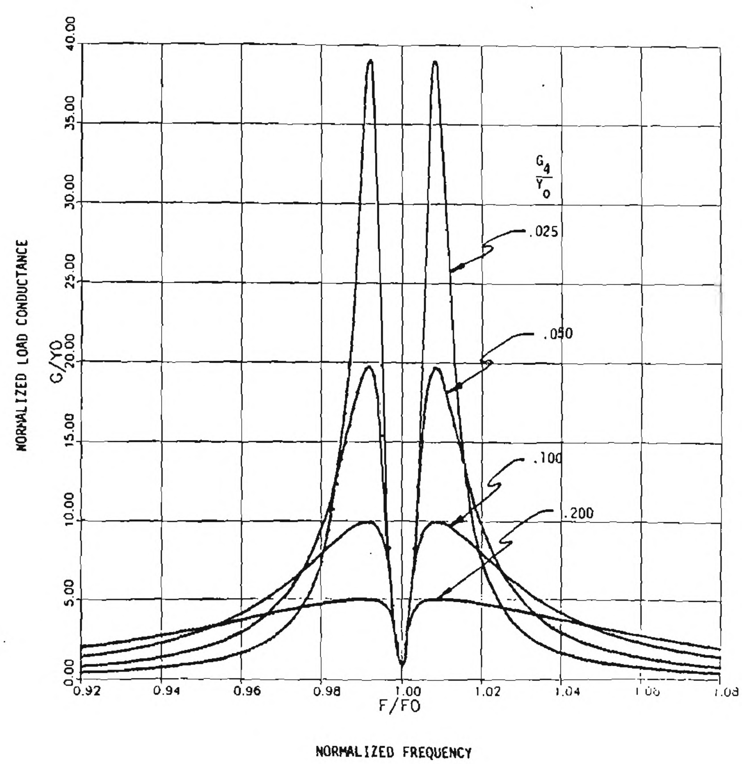


Figure 6-14. Load Conductance.

with the negative of the device admittance line can be studied. For the high Q cases being considered, such plots are not helpful because of the multiply resonant properties of the load.

NOTE 2): Many variations on the stabilization geometry discussed above have been demonstrated. At microwave and millimeter frequencies, every such oscillator, (using a low Q oscillator and high Q resonator) is subject to a similar set of stability criteria and to similar limits on the maximum S-factor achievable.

To achieve maximum stabilization with minimum difficulty, the following steps are helpful in every case of which we are aware.

1) Fabricate the basic oscillator with the highest possible Q before attempting stabilization.

2) If high Q per se is not possible, another means should be sought to restrict the range of frequency over which the basic oscillator exhibits a negative conductance. For two port oscillator devices, this can sometimes be accomplished in the feedback loop or at the port which is not connected to the stabilizer. Even in this case the question is one of increased Q in either the feedback or the unconnected device port.

3) The basic oscillator should be designed to deliver maximum power significantly exceeding the stable output required. A factor of two of three is advisable so that $P_3 > 0$ and $G_4 >$

0 are permitted.

4) If possible, the overall network should allow for significant adjustment so that:

   a) the basic oscillator and stabilizer can be synchronously tuned,

   b) The relative values of β_1 and β_2 can be optimized and

   c) so that G_4 may be set for minimum absorption yet large enough to obtain the required stabilization.

6.5 Circuit Realization

A possible planar circuit realization of the approach just discussed is given in figure 6-15 .

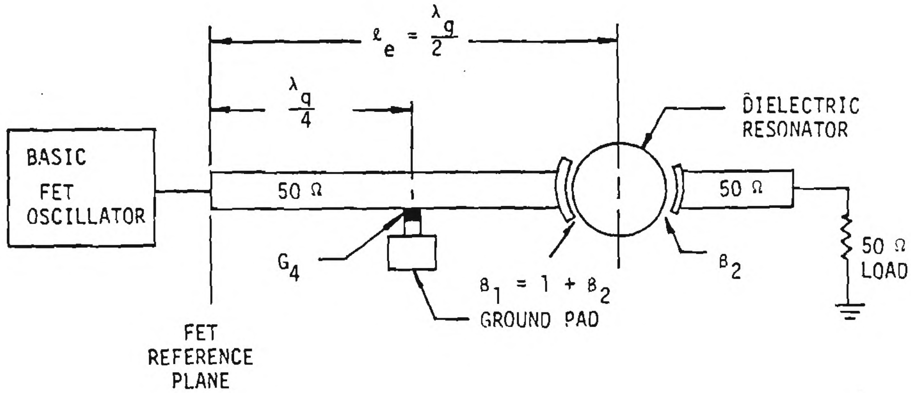
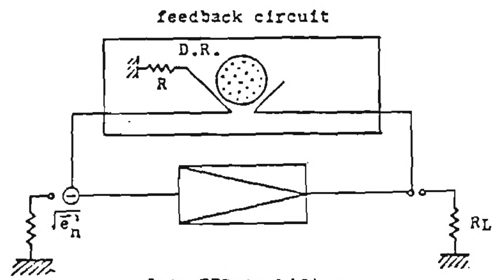
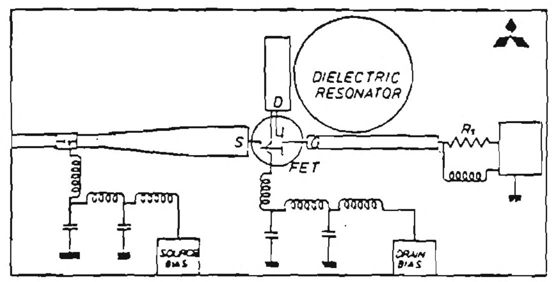


Fig. 6-15 . Possible Planar Oscillator with Dielectric Resonator, adjustable coupling (β_1, β_2) and damping (G_4).

Numerous other geometries have been reported and analyzed. The feedback approach of Mori(6-9) (figure 6-16) is of interest as are those of Tsironis(6-10) and Podcameni(6-11) shown in figures 6-17 and 6-18 . These three oscillators were not designed for millimeter frequencies. Some additional approaches with more direct application to millimeter frequencies are given in references (6-12) through (6-16).



Block diagram of a DRF GaAs FET oscillator circuit



MIC pattern of the DRF GaAs FET oscillator

Figure 6-16 . Mori's Feedback Stabilized Oscillator.

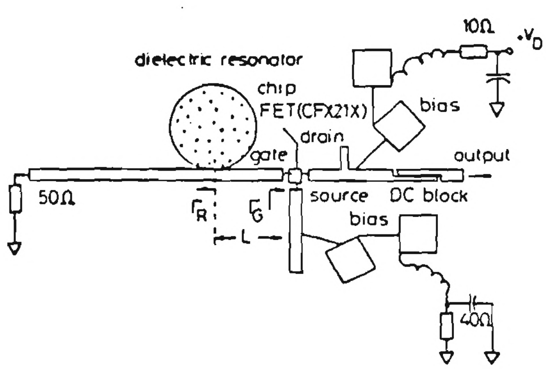


Figure 6-17 . Tsironis' Stabilized Oscillator

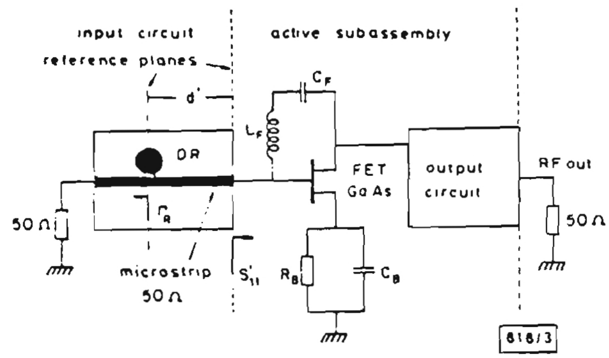


Figure 6-18 . Podcameni's Stabilized Oscillator

VII. CONCLUSIONS/RECOMMENDATIONS

7.1 Overall Receiver

The development of Ka-band monolithic receivers should proceed in much the same way as the development of 12 GHz receivers described herein. Each building block (local oscillator, IF amplifier, and mixer) should be developed and tested separately with an emphasis on compatibility with respect to interface connections, impedance matching and fabrication techniques.

The first stages of monolithic mixer development should be performed using a low frequency model where measurements are made more readily and where empirical adjustments to the circuits are more easily performed. The circuits can then be scaled accordingly.[7-1, 7-2]

Each mixer type for which approximate designs are discussed has it own advantage and disadvantages. Each of these could be developed in parallel or one particular mixer can be pursued based on the results presented here. The best mixer receiver performance would probably result from the crossbar mixer. The highest yield would likely come from the balanced Lange or direct coupled hybrid mixer or the single ended mixer. The most desirable in developing a fully monolithic receiver in the near term would result from the subharmonic mixer receiver because of the lower LO frequency requirements. A summary of these receivers is given in Table 7-1.

7.2 Local Oscillator

Half micron FET geometries allow use of the FET as either a half frequency subharmonic pump or a half frequency voltage tuned doubling oscillator. A separate doubler is also feasible. Addition of either active stabilizing circuits (locked loop, discriminator) or passive elements leads to overall circuit dimensions on GaAs that range up to 0.5 inch Therefore, initial design and or fabrication work on the local oscillator will probably need to be done using a separate GaAs substrate. More compact designs would then be developed.

Useful information related to both passive and active stabilization can be acquired using vertical geometry Gunn device chips mounted adjacent to the remainder of the L.O. system (e.g. planar varactors, planar detectors, discriminator circuit, dielectric resonators). If 35 GHz FETs are unavailable and a subharmonic FET oscillator is deemed unsuitable, this approach could prove useful.

A critical problem related to any local oscillator approach is the adjustment of the circuit to compensate for both device and circuit tolerances. A patent solution to this problem has not been reported. Therefore, considerable initial effort must be devoted to this area.

TABLE 7-1

Summary of Monolithic Receivers

Mixer Receiver Type	Estimated Performance (SSB Noise Figure bandwidth)	Comments
RF Amplifier	4.0 dB/10% 10 dB/26.5-40 GHz	Requires 1/4μm gate FET
Balanced Hybrid	8.5 dB/36-39 GH	High yield, Airbridge maybe required, Dual Diodes
Balanced Crossbar	7.5 dB/30-32 GHz 9.0 dB/31-39 GHz	Low Yield, Inline IF/LO/RF, Requires RF Wave-guide Input Dual Diodes
Subharmonic	8.5 dB/36-39 GHz	15 GHz LO Dual Diodes, Low Yield
Single Ended	9.0 dB/36-39 GHz	High Yield Single Diode Low Risk

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